TRANSISTOR SINEWAVE OSCILLATORS

ROBERT J. GREENWAY

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Robert J. Greenway

TRANSISTOR SINEWAVE OSCILLATORS

by

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Captain, United States Marine Corps

Submitted in partial fulfillment
of the requirements
for the degree of
MASTER OF SCIENCE
IN
ENGINEERING ELECTRONICS

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MASTER OF SCIENCE

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PREFACE

The transistor, so recently invented, came into being with many applications particularly suited to it. Thus, it has been deprived of the long developmental period afforded the vacuum tube, which grew along with the advance of the radio art to become the versatile heart of so many of the complex electronic devices in use today.

The demands of industry require that this new device be put to work immediately - before the technology of fabricating it is fully developed and before its functioning is fully understood. As a result, much of the engineering in this field is approached from an empirical point of view with only general guidance from a rather complex and incomplete theory of operation.

While much progress is being made, it appears that the day when transistor circuitry will have achieved the handbook stage now enjoyed by the vacuum tube is still rather far in the future.

Most of the developmental work discussed in Chapter V was accomplished while working with the Lenkurt Electric Company, San Carlos, California under the industrial tour program of the U. S. Naval Postgraduate School.

The assistance and suggestions of Mr. B. R. Stack, Project Engineer of that company, were indispensable in this work and the writer wishes to express his deepest

appreciation.

The patience and guidance of Professor G. R. Giet, Chairman, Electronics Department, U. S. Naval Postgraduate School, were a major contribution to this paper and the writer wishes to extend his most grateful thanks.

Thanks are also due the Lenkurt Electric Company for the reproduction of many of the illustrations included in this thesis.

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TABLE OF SYMBOLS AND ABBREVIATIONS

r_i	- Input resistance
r_o	- Output resistance
r_b	- Internal base resistance
r_c	- Internal collector resistance
r_e	- Internal emitter resistance
R_g	- Resistance across an input circuit
R_L	- Resistance across an output circuit
r_m	- Common-base transfer resistance
R	- A resistance in general
L	- An inductance in general
C	- A capacitance in general
N	- Transformer turns ratio
A_i	- Current amplification
v_c	- A-c. collector voltage
R_{eff}	- Effective resistance
i_c	- A-c. collector current
i_b	- A-c. base current
g_m	- Transconductance
V_c	- D-c. collector voltage
I_c	- D-c. collector current
I_b	- D-c. base current
α	- Alpha; short-circuit current amplification factor
$\overline{g_m}$	- Average transconductance

I_{co}	- D-c. collector current for zero emitter current; frequently referred to as "collector saturation current"
V_{cc}	- Power supply voltage
b	- Common-emitter current amplification factor
db	- Decibel(s)
$f_{\alpha o}$	- Frequency at which $\lvert\alpha\rvert$ is reduced by 3 db
f_{bo}	- Frequency at which $\lvert b \rvert$ is reduced by 3 db
X	- Reactance; imaginary part of an impedance
$^{o}C.$	- Degree(s) centigrade
ma.	- Milliamperes
mw.	- Milliwatt(s)
rms.	- Root-mean-square
RMA	- Radio Manufacturers Association
mh.	- Millihenry(ies)
kcs.	- Kilocycles per second
mcs.	- Megacycles per second
I_e	- D-c. emitter current
P_{max}	- Maximum power dissipation
p	- Stabilization factor

CHAPTER I

INTRODUCTION

1. General.

The engineer confronted with the necessity of designing an oscillator employing a transistor in lieu of a vacuum tube has at his disposal a rather large, and constantly growing, body of literature. However, much, if not most, of this literature is found to bear upon cases of very limited application. The writer was able to find but a single case where an oscillator had been designed for quantity production [1] and this was made possible by special arrangements with the transistor manufacturer who carefully selected and color coded transistors suitable for the specific circuit.

On the other hand, information is available which presents a general analysis of transistor oscillators. Here, the variations of the parameters from transistor to transistor or for a given transistor in a changing environment are usually assumed nonexistent.

Between the special case of limited usefulness and the general analytic solution lies the area of greatest interest from an engineering point of view. It is this area which must be developed if the transistor is to achieve the promise which has been so widely claimed for it.

2. Summary.

This paper will deal almost entirely with this middle

ground. Several of the more promising circuits will be presented with a discussion of their more important advantages and disadvantages. Where appropriate, comparisons will be drawn with vacuum tube circuits to point up shortcomings or strong points of the transistor. In addition, the detailed steps in the design and development of a practical oscillator for quantity production will be presented to illustrate one approach to the problem. The results of performance tests will be compared with the design specifications to indicate the extent to which the final circuit fulfills the requirements.

CHAPTER II

BASIC TRANSISTOR OSCILLATOR CIRCUITS

1. General.

Considered from the broad point of view, the same two groups of oscillators available in vacuum-tube circuits are also possible in conjunction with transistors: negative-resistance oscillators and feedback oscillators. These descriptive titles maintain the same meanings in transistor work that is associated with them in tube circuits. The negative-resistance oscillator is a two-terminal network in which the energy dissipated in lossy elements is balanced by energy introduced by the negative-resistance elements. The feedback oscillator consists of an amplifier with a gain greater than unity in which sufficient energy to overcome the circuit losses is fed back in the proper phase from the output to the input terminals.

2. Negative-resistance Oscillators.

With transistors having an alpha greater than unity, it is possible to make the real part of the impedance looking into either the output or input of any of the three standard configurations appear negative [2]. By "three standard configurations" is meant the common-base, common-emitter and common-collector circuits. The requirement that alpha exceed unity precludes the use of junction tran-

sistors as negative-resistance oscillators. Hence it should be understood that the discussion of negative-resistance oscillators which follows applies only to the point-contact device.

Liben [2] shows that for the common-base circuit as shown in Figure 1 the input resistance, r_i, or the output resistance, r_o, will become negative and spontaneous oscillations result when the inequality

$$Z_t + r_b > \frac{(R_g + r_e)(R_L + r_c)}{r_m - (R_g + r_e + R_L + r_c)} \qquad (1)$$

is satisfied. Whether r_i or r_o becomes negative depends upon the relative values of R_g and R_L. Refer to Appendix II for the derivation of inequality (1).

Letting Z_t take the form of a parallel resonant circuit as in Figure 2,

$$Z_t = \frac{\frac{R}{(\omega C)^2} - j\left[\frac{R^2}{\omega C} + \frac{L}{C}\left(\omega L - \frac{1}{\omega C}\right)\right]}{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}$$

The magnitude of Z_t will be large when $\omega^2 = \frac{1}{LC}$ and small for other values of ω. Thus, inequality (1) is satisfied and oscillations will result at the resonant frequency of Z_t.

For the common-emitter circuit, oscillations occur when

$$R_L < \frac{(r_m - r_e)(R_g + r_b)}{R_g + r_e + r_b} - r_c \qquad (2)$$

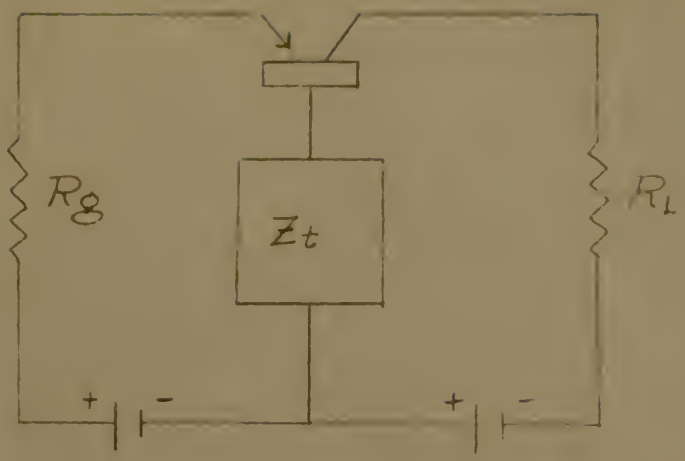


Figure 1

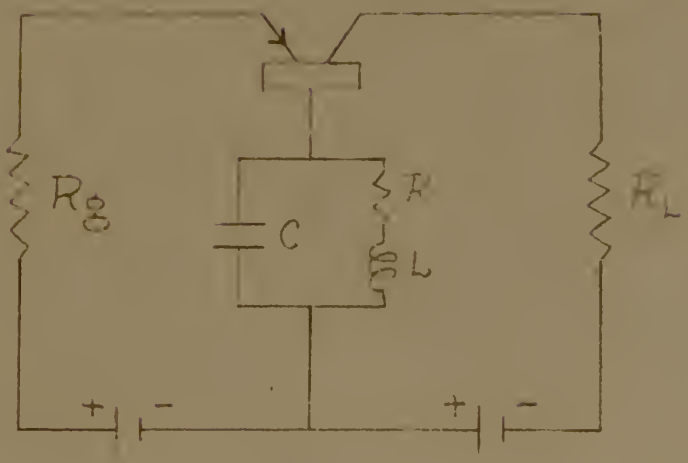


Figure 2

As shown in Figure 3, R_L, the parallel combination of Z and RLC, becomes small at the resonant frequency of the series RLC circuit and oscillations will occur. Z is merely a high a-c. impedance to provide a means of applying d-c. bias to the collector.

If desired, the series RLC circuit may be replaced by a suitable quartz crystal for crystal-controlled frequency stability.

Solving inequality (2) for r_e yields

$$r_e < \frac{(R_g + r_b)(r_m - r_c - R_L)}{R_L + r_c + R_g + r_b} . \qquad (3)$$

Since r_e is positive, $r_m > r_c + R_L$ must exist for (3) to be satisfied. These conditions lead to the common-emitter circuit shown in Figure 4. For oscillations to occur, the impedance of the parallel circuit in the emitter branch added to r_e must be smaller than the right-hand member of (3).

Inequality (2) may also be solved for R_g, giving

$$R_g > \frac{r_e (r_m - r_c)}{R_L + r_e + r_c - r_m} - (r_e + r_b) \qquad (4)$$

This condition may be achieved by replacing R_g by a parallel resonant circuit which will present a large resistance at its resonant frequency. This circuit is illustrated in Figure 5.

The grounded-collector circuit becomes oscillatory if

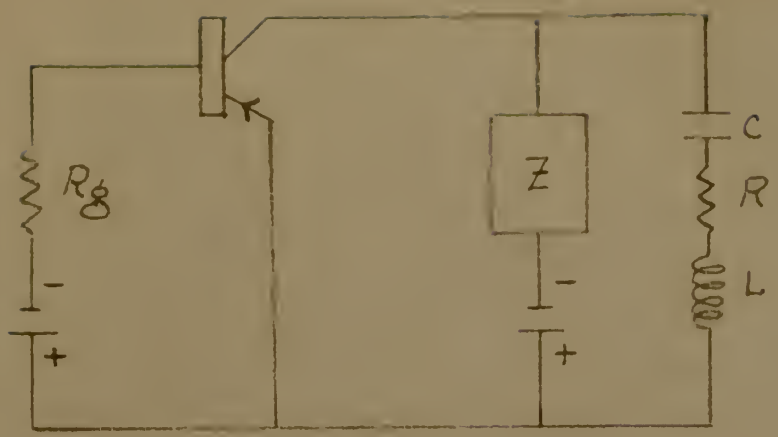


Figure 3

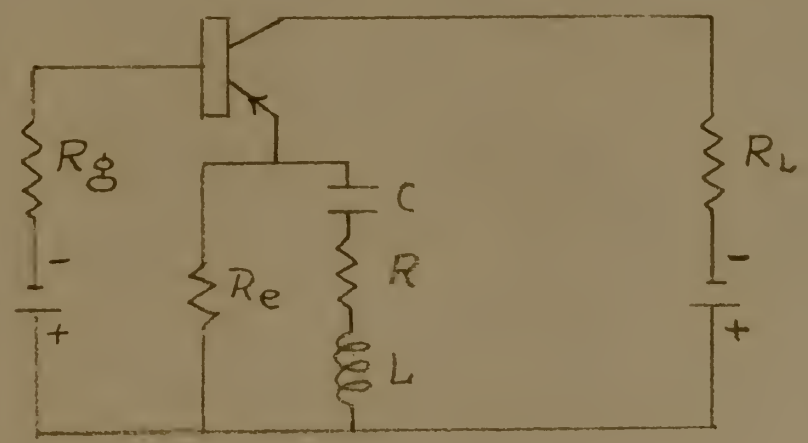


Figure 4

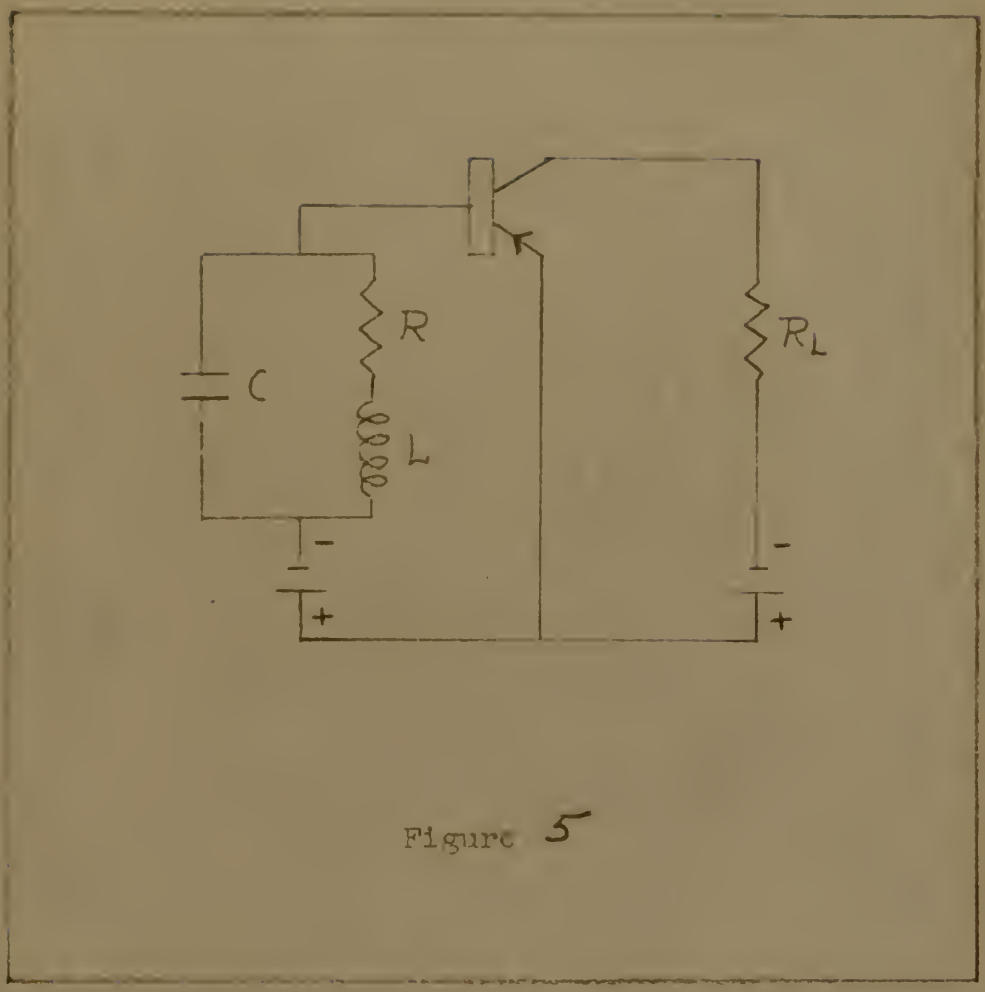


Figure 5

$$R_g > \frac{r_c (r_c - r_m)}{R_L + r_e + r_c - r_m} - (r_c + r_b) \qquad (5)$$

or

$$R_L < \frac{(R_g + r_b)(r_m - r_c)}{R_g + r_c + r_b} - r_e \qquad (6)$$

Inequality (5) leads to the circuit of Figure 6 and (6) may be represented as in Figure 7.

The circuits shown do not represent all possible configurations of negative-resistance oscillators. Other circuits are readily obtained by choosing circuit parameters such that the equations for the input or output resistance yield a negative result.

3. Feed-back Oscillators.

As stated earlier, this type of oscillator consists essentially of an amplifier. As such, any of the three standard configurations again may be employed. There is no restriction imposed in this case that the transistor be a point-contact device. However, since the current gain should exceed unity, if the grounded-base configuration is used a transformer will be required when junction transistors are employed since they have an alpha less than one.

The block diagram of the transistor feedback oscillator is identical with its vacuum-tube counterpart, as may be seen from Figure 8. However, it is generally more convenient and more meaningful to work in terms of current in

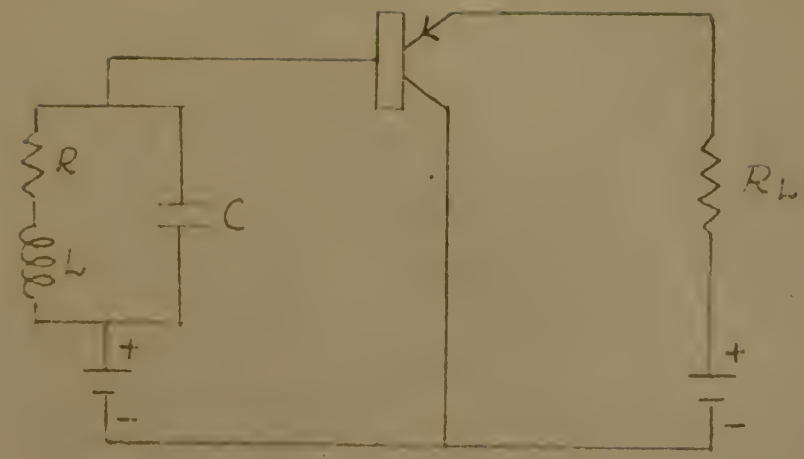


Figure 6

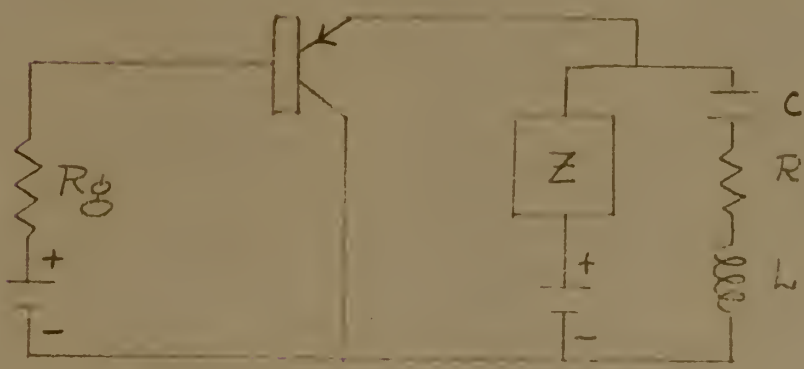


Figure 7

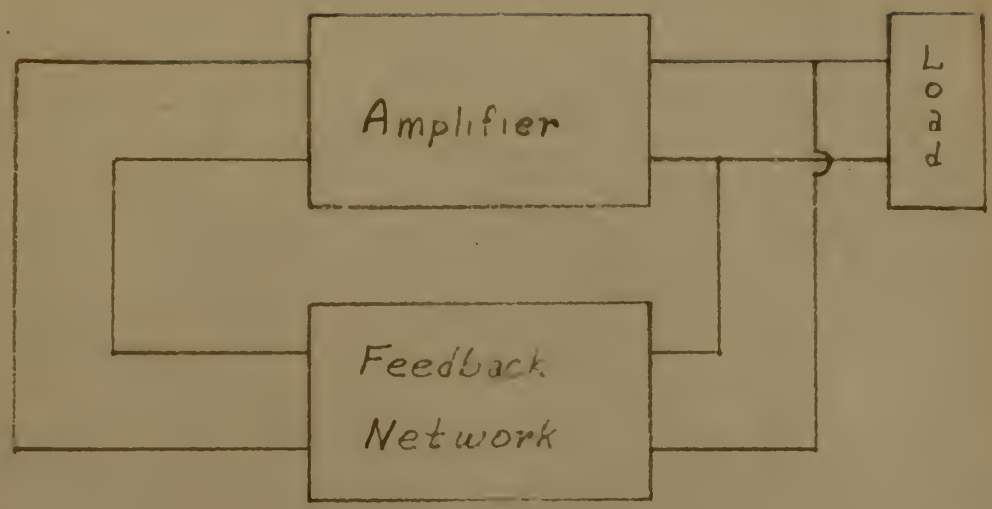


Figure 8

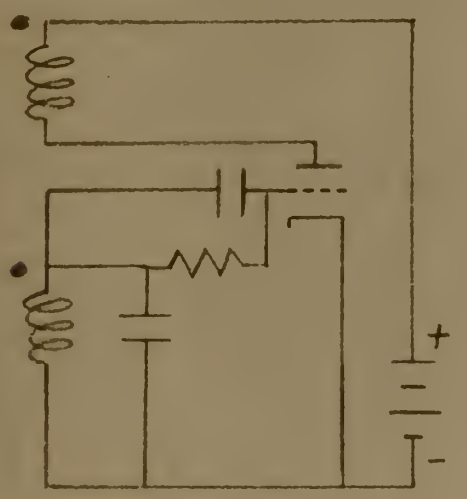


Figure 9

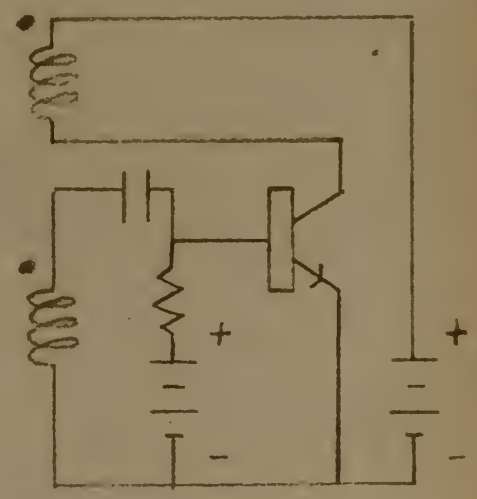


Figure 10

in connection with transistors. This is true because in some cases current is a multi-valued function of voltage while voltage is single-valued with respect to current. This is the dual of the situation occurring in the vacuum tube tetrode plate characteristics where a negative resistance is exhibited. Consequently, feedback will be considered in terms of current.

The circuit frequently employed in introducing vacuum-tube feedback oscillators is shown in Figure 9. A circuit which will serve a similar purpose in transistor work is that of Figure 10. While it is not intended that a detailed and general analysis of this circuit be presented, the literature is singularly devoid of any discussion of the means by which the amplitude stability is achieved in transistor oscillators and perhaps a brief discussion of this facet is justified. For this, the circuit of Figure 10 will serve our purpose.

The coefficient of coupling of the transformer in the circuit is sufficiently near unity that this may be assumed to a very close approximation. The current induced in the base-emitter circuit is then -N times the collector current, where N is the ratio of collector winding turns to the base-emitter winding turns, or

$$i_b = -Ni_c \qquad (7)$$

the negative sign arising from the phase reversal required

to compensate for the phase reversal experienced by the current through the common-emitter configuration.

The current gain for this circuit is

$$A_i \approx -\frac{\alpha}{1-\alpha} = \frac{i_c}{i_b} \qquad (8)$$

or

$$i_b = -i_c \frac{(1-\alpha)}{\alpha} \qquad (9)$$

Comparing (9) with (7) indicates that for oscillations to be sustained,

$$N \geq \frac{1-\alpha}{\alpha} \qquad (10)$$

must be satisfied.

The frequency of oscillation is largely determined by the series resonant frequency of the base-emitter circuit.

As in the case of the vacuum-tube oscillator, a non-linearity is responsible for the stable operation of the transistor oscillator. This will require a departure from small-signal theory and point up the reason that these oscillators are inherently incapable of producing a pure sinewave output, free of harmonics.

By drawing a load line on the I_c-V_c characteristics for the common-emitter configuration, it is possible to construct a characteristic curve such as that shown in Figure 12. The output characteristics for the TI-200 will be found

Transfer Characteristic

6SJ7

$R_L = 18K\Omega$

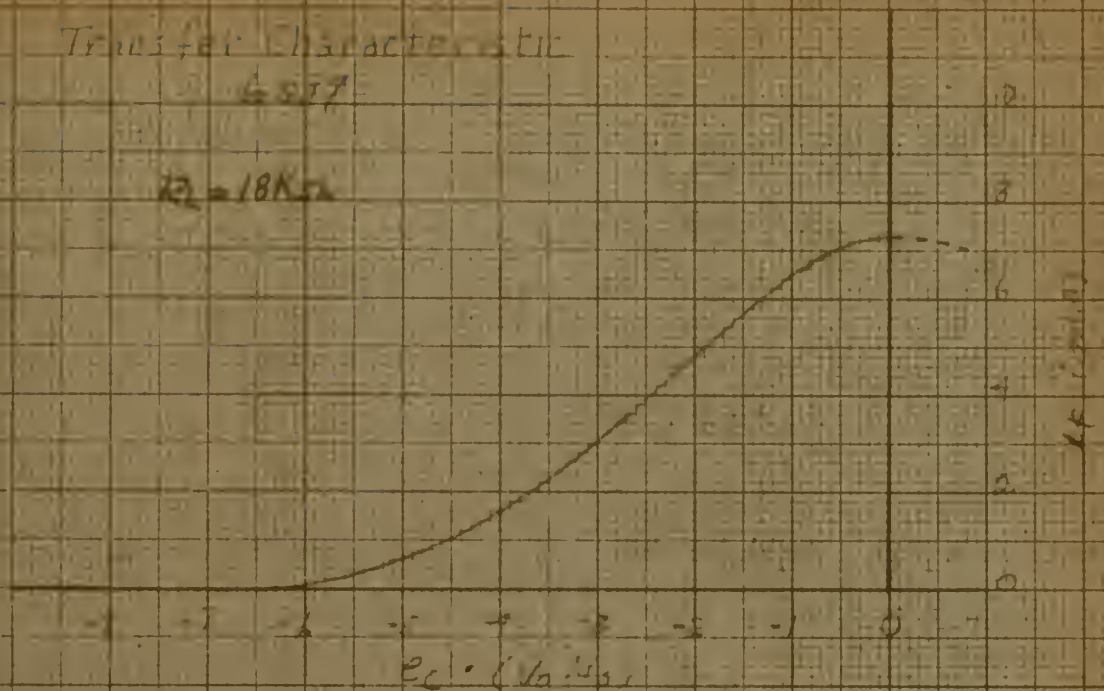


Figure 11

Transfer Characteristic

TT-302

$R_L = 6K\Omega$

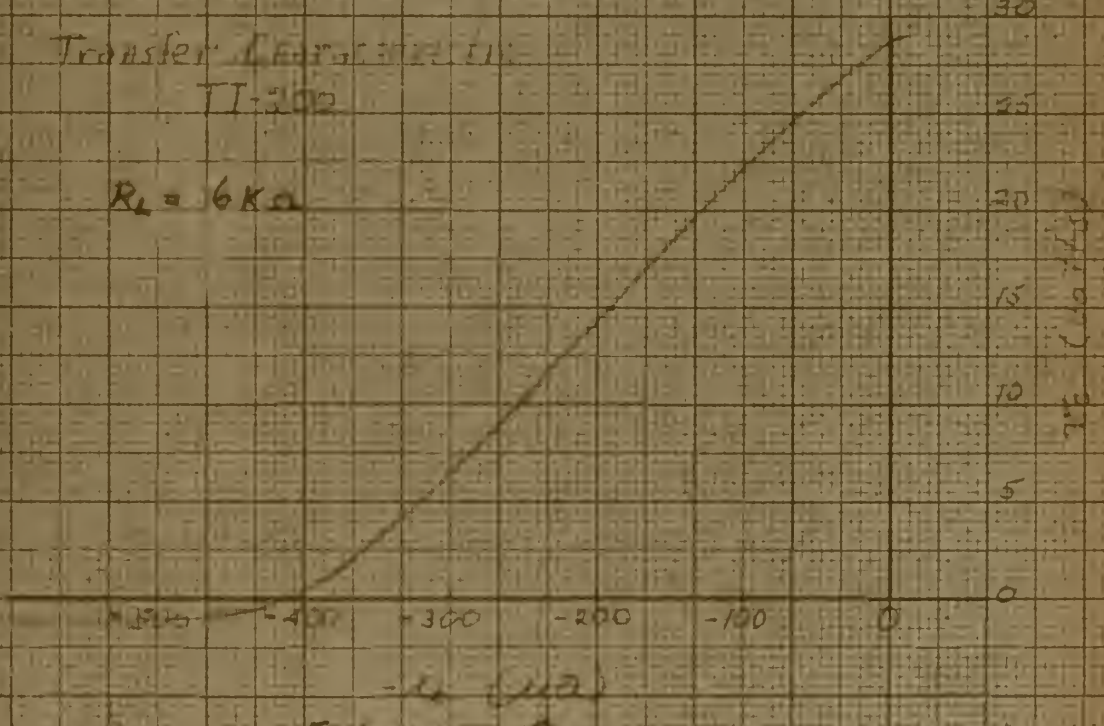


Figure 12

in Appendix III. The load used in this illustration is 6000 ohms. For comparison, the transfer characteristics of a 6SJ7 with recommended biasing is shown in Figure 11.

The slope of the vacuum-tube transfer characteristic has the dimension of comductance and is the generally recognized transconductance, g_m. In the case of the transistor, the slope of the curve has the dimension of resistance and is the equivalent base-collector transfer resistance. Here one must be careful to delineate clearly between the emitter-collector transresistance, r_m, and the resistance just described.

The a-c. collector voltage

$$v_c = i_c R_{eff} \qquad (11)$$

where R_{eff} is the effective load resistance seen by the collector.

From (8), $$i_c \approx -i_b \left(\frac{\alpha}{1-\alpha}\right)$$

Substituting,

$$v_c \approx -i_b \left(\frac{\alpha}{1-\alpha}\right) R_{eff} \qquad (12)$$

or

$$\frac{v_c}{-i_b} \approx \left(\frac{\alpha}{1-\alpha}\right) R_{eff} \qquad (13)$$

which is seen to be the slope of the transfer characteristic

of Figure 12.

The characteristic shown was obtain from a 6000-ohm load line through the operating point V_c = 15 volts and I_b = 190 microamperes, from the common-emitter output characteristics of a TI-200 transistor.

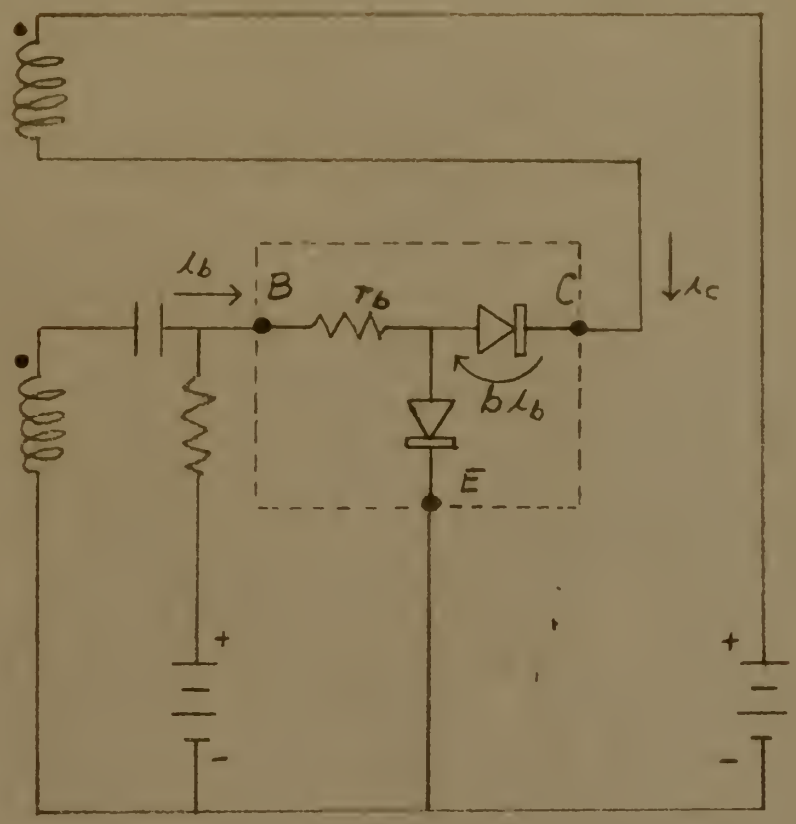
This transistor has an alpha varying from 0.90 to 0.95. Assuming an average value of 0.925 and substituting in (13) gives a slope of 74,000 ohms. On the transfer characteristic, swinging 100 microamperes either side of the operating point gives a slope of 77,500 ohms which is reasonable agreement.

For the vacuum-tube oscillator shown in Figure 9, it will be recalled that the amplitude of the grid voltage swing continues to grow until a value of $\overline{g_m}$, an average g_m, is reached that just sustains oscillations.

Similarly, the excursions of i_b in the transistor circuit increase until an average value of the transfer resistance is reached that exactly maintains oscillations.

The mechanism by which the amplitude of the vacuum-tube oscillator output is stabilized is well known and will not be discussed further. A few words may be in order regarding the transistor circuit, however.

The slope of the transistor transfer characteristic shown is much more nearly constant in its center range than that of the vacuum tube curve. Also, its nonlinearities are



Large Signal

Equivalent Circuit

Figure 13

much more promounced. The reasons for this become evident when the large-signal equivalent circuit is observed. In Figure 13, the equivalent circuit is shown for the large-signal operation of an NPN transistor. The base-emitter diode is biased in the forward direction such that it is effectively a low resistance quiescently. The base-collector diode, however, is biased in the reverse direction and is a high impedance quiescently.

With large alternating components of current through these diodes, it is possible for either or both of them to have the current in them reduce to zero and reverse directions. This results in the abrupt nonlinearities in the transfer characteristic.

Suppose that the collector circuit were to induce into the base circuit a current sufficiently large so that, on its negative swing, it tended to reverse the direction of i_b from that shown in Figure 13. The base-emitter diode would then appear as a high impedance (the backward impedance of the diode) and essentially no current would flow. At this point, the transistor ceases to behave as such and the current flow in the collector circuit is reduced to I_{co} with the result that V_c rises to and remains at a voltage equal to $V_{cc} - I_{co}R_L$, where R_L is the effective resistance in the collector circuit. This condition will persist until the Zener breakdown is reached in the base-emitter diode or

until the diode returns to forward conduction.

At the opposite end of the transfer characteristic, the base-collector diode brings about the nonlinearity. This occurs when the current swing in this diode is of such a magnitude to cause the diode to conduct in its forward direction. In this condition, it becomes a low resistance and the collector drops to a small negative potential due to the drop across the base-collector diode. This small negative potential remains nearly constant for relatively large swings in i_b.

These two nonlinearities may both exist, on opposite half-cycles, if the operating point is chosen near the center of the linear portion of the transfer characteristic.

The excursions of i_b beyond the linear portion of the transfer characteristic cause the average transfer resistance to be reduced just as $\overline{g_m}$ was reduced in the case of the vacuum tube.

A value of this average transfer resistance will eventually be reached which will result in stable operation. As in the case of the vacuum tube, the slope of the characteristic in the vicinity of the operating point must be sufficiently steep to produce growing oscillations if the oscillator is to be self-starting.

The fact that the nonlinearities in the curve are rather abrupt leads to hard clipping. It is evident that this re-

sults in considerable harmonic content being present in the output of the circuit.

Two other feedback oscillator circuits are shown in Figures 14 and 15.

Figure 14 represents a common-base circuit employing a point-contact transistor. Shea [3] shows that the feedback current ratio is approximately equal to 1/a. Since a is approximately equal to alpha, the current feedback ratio exceeds unity for an alpha less than one, as in the junction transistor. Hence the necessity for a point-contact device in this circuit. He further shows the conditions for oscillation are

$$C_2 = C_1\left(\frac{a}{1-a}\right) \qquad (14)$$

and

$$\omega^2 = \frac{C_1 + C_2}{L(C_1C_2)} \qquad (15)$$

with (15) defining the frequency of oscillation. This circuit bears a strong resemblance to the Colpitts circuit with L bridging C_1 and C_2.

The circuit of Figure 15 requires a feedback ratio of at least 1/b where b is the common-emitter current amplification factor. Since b exceeds unity for both junction and point-contact transistors, either may be used here.

It is readily shown that the conditions for oscillation in this configuration are

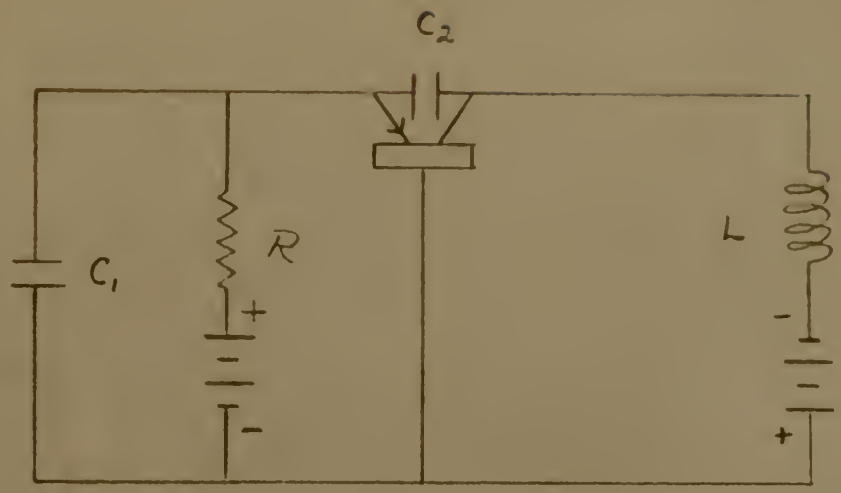


Figure 14

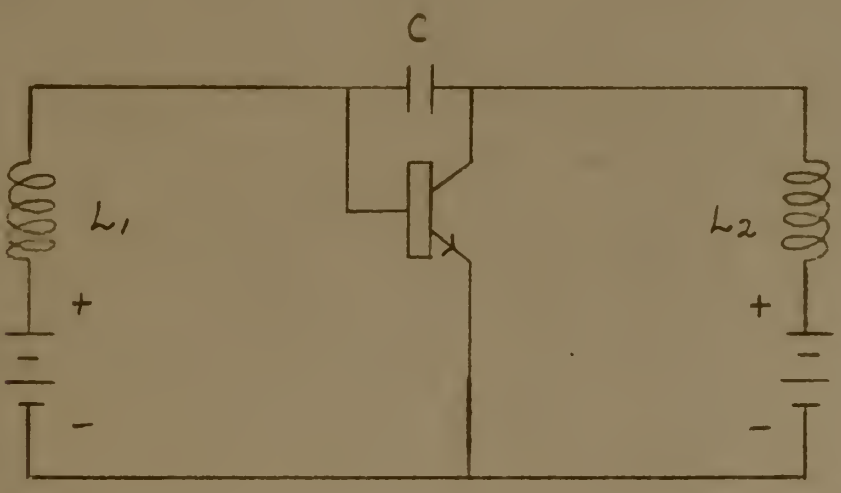


Figure 15

$$L_2 = L_1\left(\frac{a}{1-a}\right) \qquad (16)$$

and

$$\omega^2 = \frac{1}{C(L_1 + L_2)} \qquad (17)$$

where (17) indicates the frequency of oscillation. This circuit resembles the Hartley circuit somewhat, with C bridging L_1 and L_2.

While relatively few feedback oscillator circuits have been illustrated, it will be observed in later chapters that these simple circuits embody the principles of what appear to be more complex circuits. As in vacuum-tube oscillators, there are many specific transistor oscillator circuits, but they are largely variations of the few basic configurations.

CHAPTER III

PRACTICAL TRANSISTOR OSCILLATORS

1. General.

In this chapter, some typical transistor oscillator circuits of a practical nature will be discussed. These have been taken from the literature and are presented to give an indication of the present state of the art.

Circuits utilizing both point-contact and junction transistors are shown. Also, both negative-resistance and feedback types will be discussed.

2. Typical Circuits.

The circuit shown in Figure 16 is a 50 megacycle oscillator [4] employing an RCA 2N33 point-contact transistor in a negative-resistance circuit. The configuration is of the common-base variety shown in Figure 2. The power to be expected from a circuit of this type is of the order of one milliwatt, with this transistor. The transistor is shunt fed through the 1.0 mh choke, with the proper bias being developed across the 1.0-K and 5.1-K resistors. The very small, 1.0 uuf, capacitor in the emitter circuit is required for phase correction at this relatively high frequency. This phase correction problem will be discussed in Chapter IV.

A circuit of the Clapp type is shown in Figure 17. This oscillator [5] employs a junction transistor, shown here

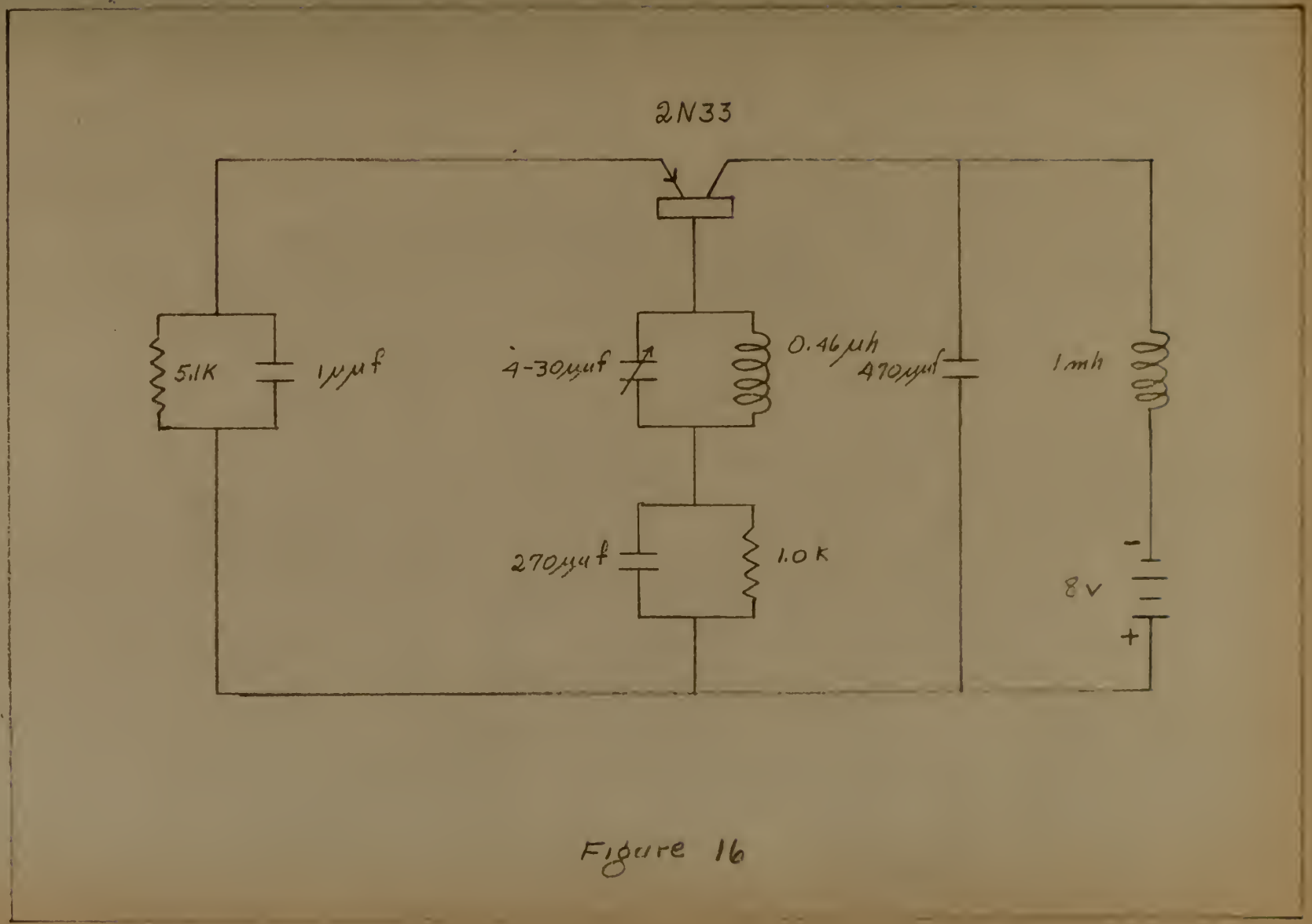
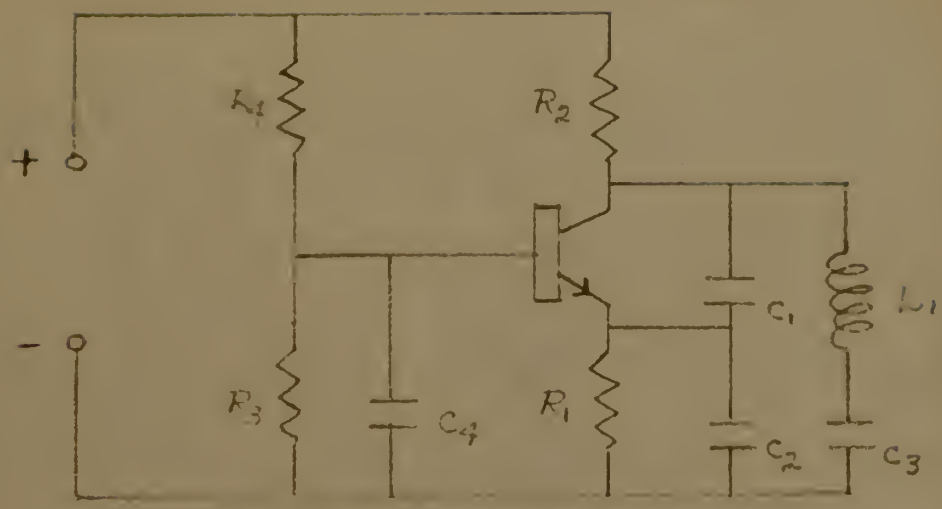
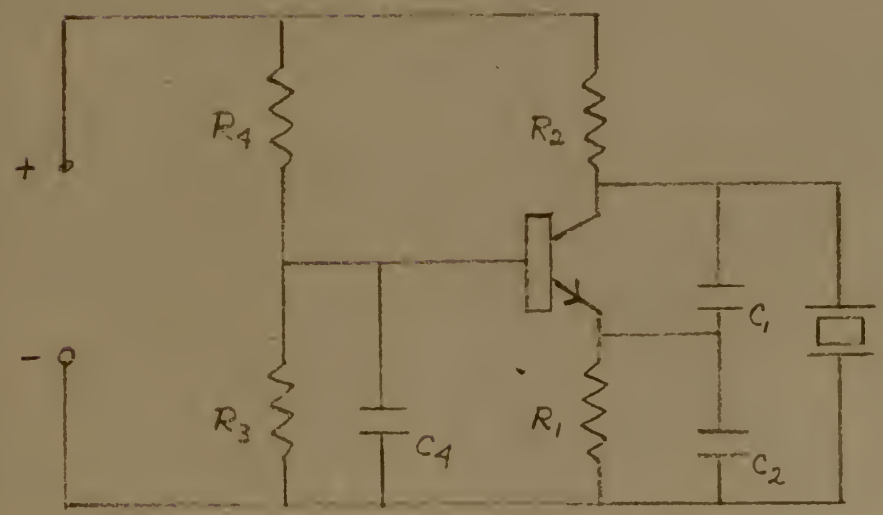


Figure 16



"Clapp" Oscillator

Figure 17



Crystal-Controlled Osc.

Figure 18

to be an NPN. This is essentially the circuit of Figure 4. However, in this case the transistor is biased with a single battery supply. A Texas Instrument Co. TI-200 functions well in the configuration, as does the Raytheon CK722 provided the polarity of the bias supply is reversed, since the CK-722 is a PNP transistor. This circuit functions well up to several megacycles with the transistors mentioned, the upper frequency limit being a function of the alpha cut-off frequency.

An interesting variation of this circuit is shown in Figure 18, with a crystal replacing L_1 and C_3. In both of these circuits, the four resistors provide for the proper biasing of the circuits, C_4 being a by-pass capacitor. The frequency is determined by C_1, C_2, C_3 and L_1 exactly as in the tube version of the Clapp circuit. Biased with a 30-volt supply and about 1.0 milliampere emitter current, a voltage coefficient of frequency of 12 cycles per megacycle per volt is possible with the circuit of Figure 17.

Figure 19 employs a Bell Telephone Laboratories M-1729 point-contact transistor in another crystal oscillator [6]. This particular oscillator was designed to supply carrier power at a frequency of 184 kilocycles in a carrier telephone system. This is a feedback oscillator which takes advantage of the fact that the common-base configuration has the a-c. collector current in phase with the emitter a-c. current. The crystal operates in its series

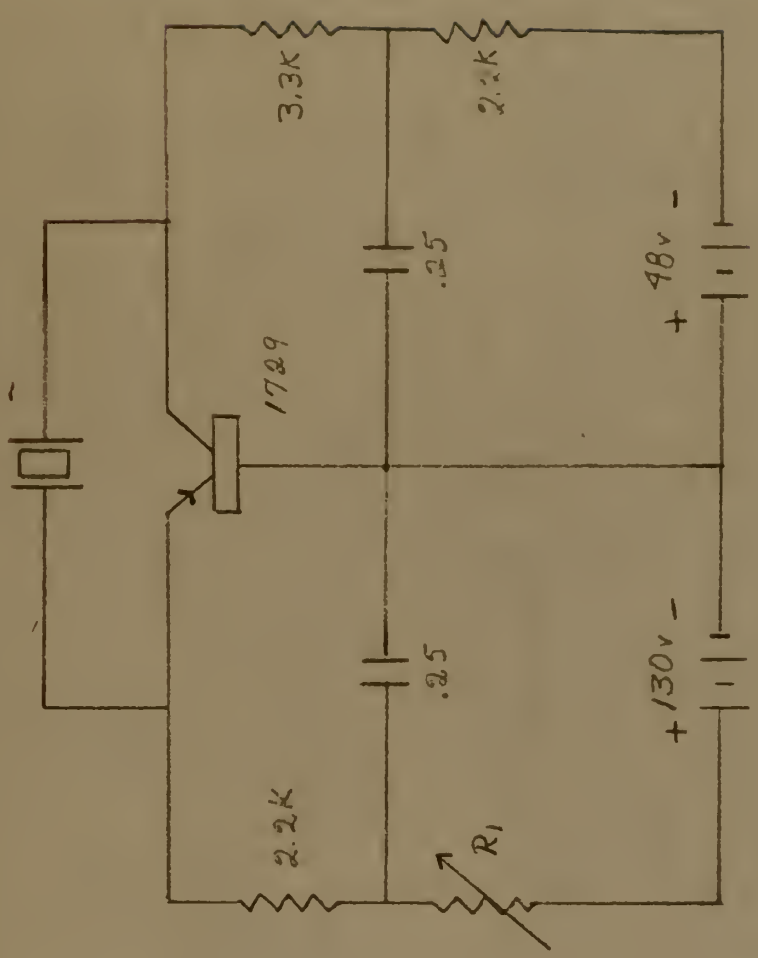


Figure 19

mode. A series resonant circuit could be employed in lieu of the crystal to produce similar results. The 130- and 48-volt supplies are standard in the carrier equipment with which the oscillator was to be employed. Resistor R_1 is adjusted for the proper value of emitter current.

The circuit of Figure 20 represents a low-power frequency standard [7]. It employs a Germanium Products Corp. RD-2517-A NPN grown-junction transistor. The crystal is a high-precision, GT-cut quartz unit. Power is supplied by a long-life, 1.30-volt mercury cell. The current drain is 100 microamperes. A potential of 0.8 volts is developed across the tank circuit. This is much too great to place across this crystal unit and the 40 uuf and 0.01 uf capacitors act as a potential divider to reduce this potential to a satisfactory level. The voltage across the crystal is coupled to the output terminal through a 100 uuf capacitor. A voltage coefficient of frequency of 0.01 cycle per megacycle per 0.10 volt is achieved with this circuit. In addition, 0.01 cycle per megacycle per degree centigrade temperature variation was noted. As compared with WWV, short-time frequency variations of plus or minus 3 parts in 10^{10} and long-time drift of 3 parts in 10^9 per 24 hours were observed.

Figure 21 shows the radio-frequency circuits of a completely transistorized, portable receiver for the standard broadcast band [1]. This includes a feedback oscillator

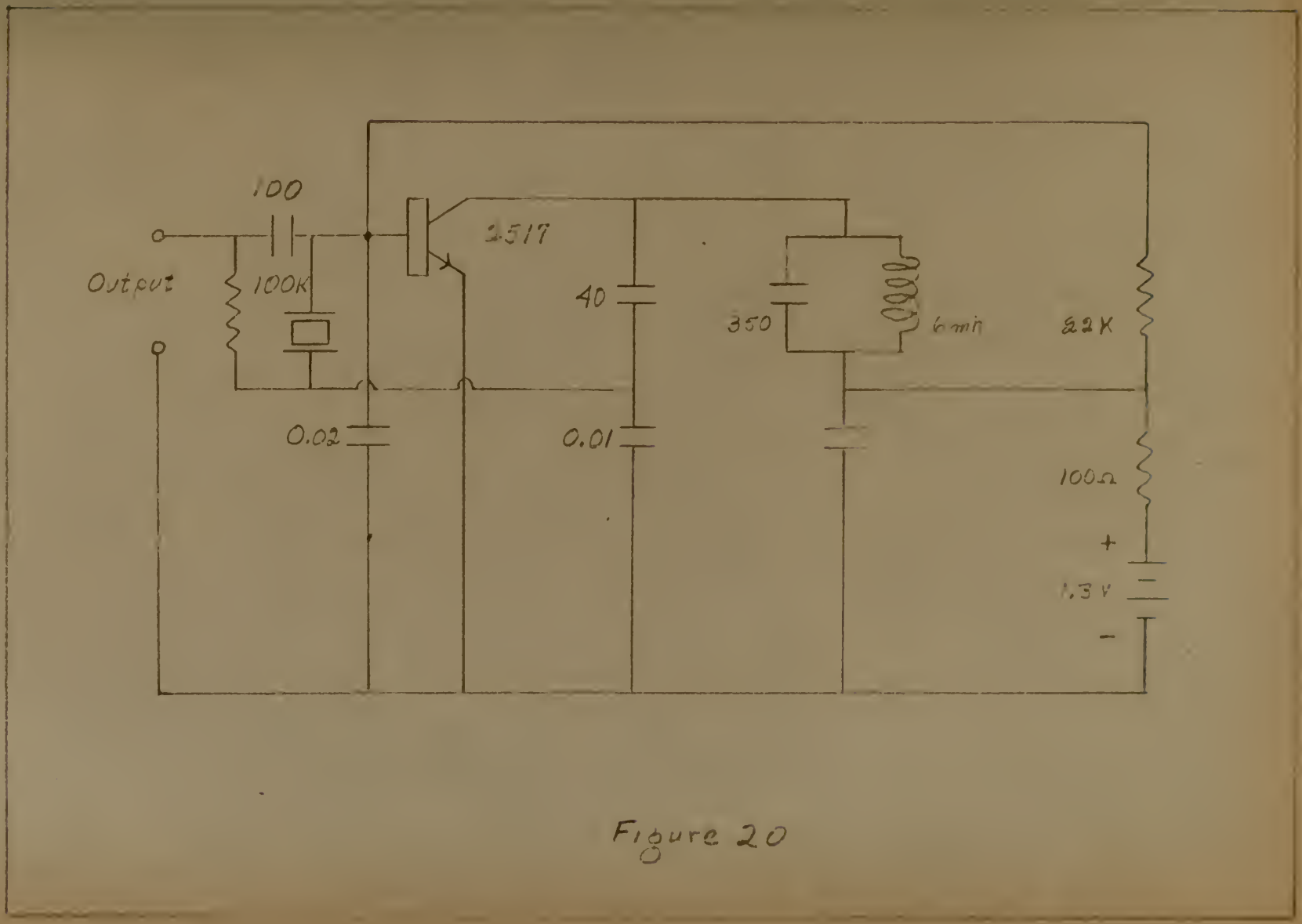


Figure 20

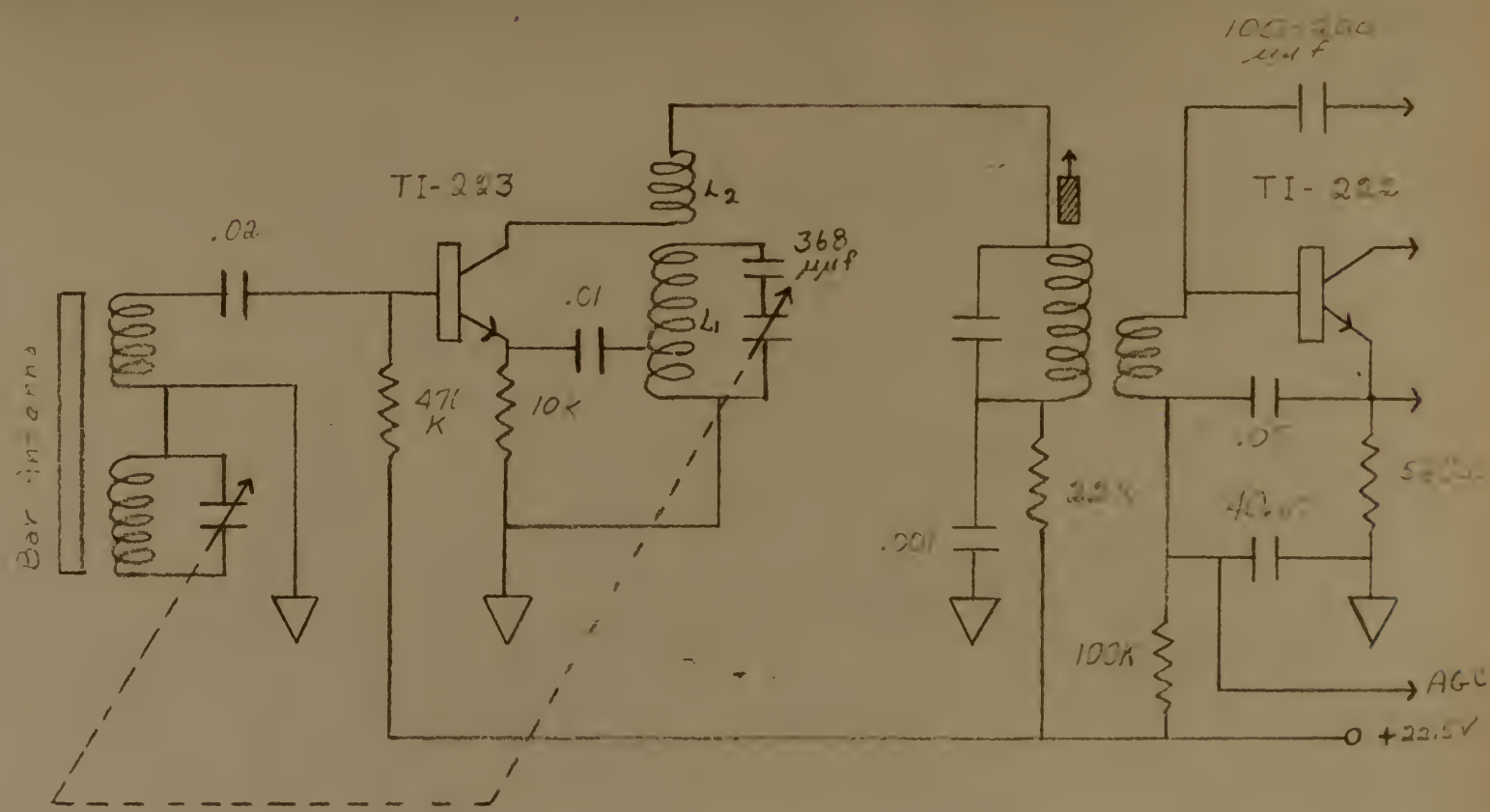


Figure 21

in which the current through L_2 induces a current in the tuned circuit containing L_1. The potential developed across this tuned circuit is large compared with the input signal level and a portion of it is coupled to the emitter. On negative swings, this oscillator voltage is sufficient to cause the "bias point" of the emitter-base diode to shift to the extent that it is reverse biased, thus introducing the nonlinearity required for mixing. The usual modulation products exist in the output in addition to the radio-frequency signal and the oscillator frequency. The tapping down on the coil is seen to keep the impedance shunting the tank circuit rather high. This is necessary in that a high Q circuit is required in order that the tuning be sufficiently sharp to prevent oscillation at the signal frequency, which is also coupled back from L_2 to L_1. The intermediate-frequency transformer performs its usual function as a band-pass filter and rejects all but the intermediate frequency. The intermediate-frequency amplifier provides 34 db power gain.

The oscillator of Figure 22, which is a complete frequency-modulated transmitter [8], utilizes a PNP junction transistor. The oscillator frequency is determined by L_1 and C_1. The feedback is via a series resonant circuit consisting of L_3 and C_3. L_2 and C_2 are tuned to the ninth harmonic of the oscillator frequency and provides an output at the frequency of 10.45 megacycles. A deviation equal to one percent of the carrier rest frequency is achieved by vary-

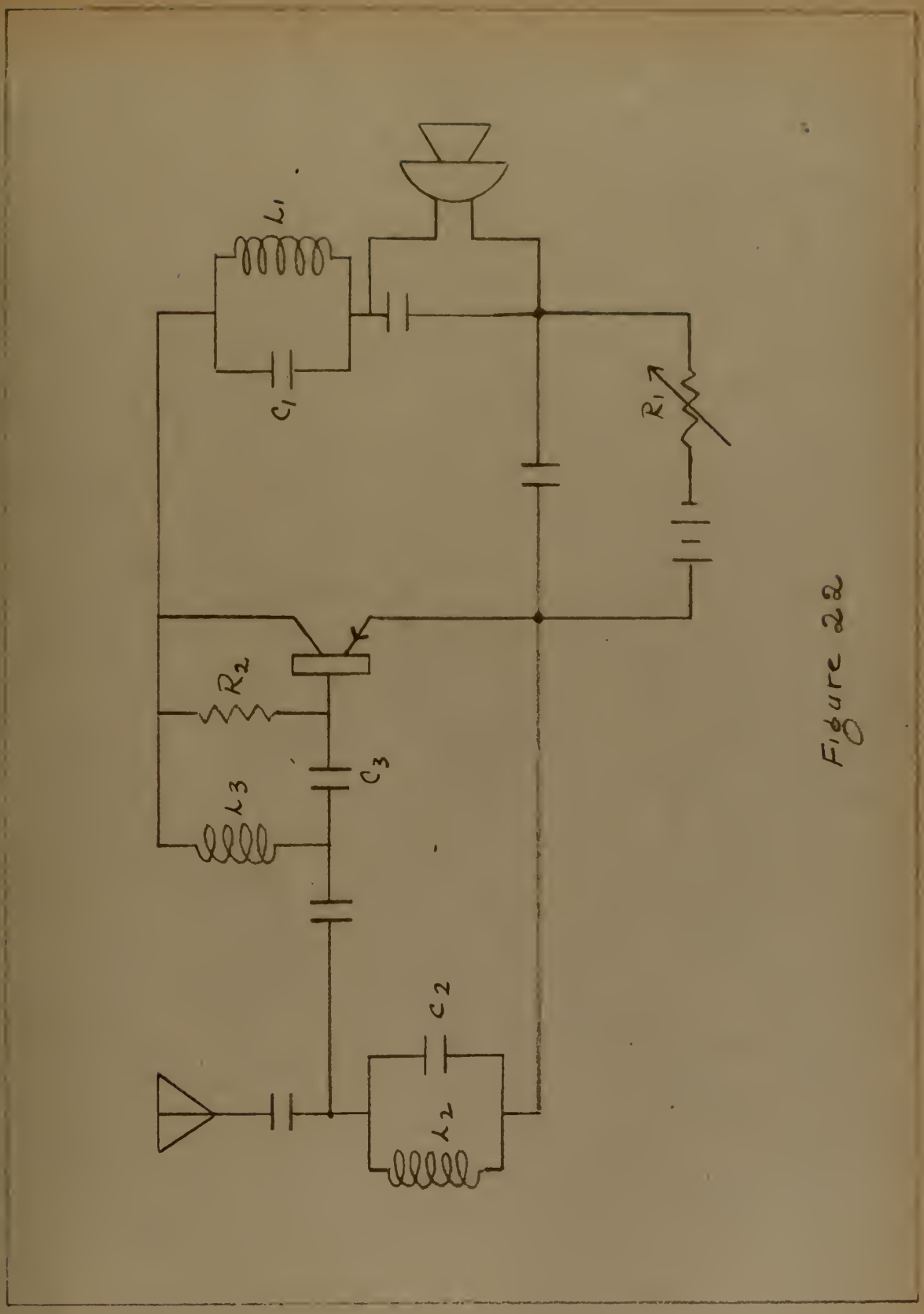


Figure 22

ing the collector voltage by one volt. This variation in collector potential is accomplished by the microphone which is in series with the supply source. The resistor R_1 is utilized for fine carrier adjustment. Amplitude modulation is produced in this circuit as a by-product but for the purpose for which the transmitter was intended, the A-M. is not considered objectionable. The transmitter was designed to be received by the intermediate-frequency amplifier of a conventional frequency-modulation receiver. By substituting a crystal in place of L_3, the F-M. may be eliminated and only A-M. produced.

3. Summary.

While it is obvious that all oscillator applications have not been discussed, it is felt that a sufficiently large cross-section has been presented to illustrate the manner in which the basic circuits illustrated in Chapter II may be implemented to provide a practical circuit for a specific purpose.

CHAPTER IV

TRANSISTOR OSCILLATOR DESIGN CONSIDERATIONS

1. General.

In approaching the design problem, the engineer will have specified for him certain performance requirements. These vary depending upon the purpose for which the oscillator is being designed. While it is not practicable to consider here all of the ramifications of the various requirements that might be imposed, there are some generalities that can be drawn. Some of these generalities will be discussed in this chapter. The application of certain of these will be illustrated in Chapter V.

2. Frequency Range.

Transistors, like vacuum tubes, have an upper bound to the range of frequencies over which they will operate satisfactorily. In general, this upper frequency is much lower for transistors than for vacuum tubes. This upper limit varies rather widely from transistor to transistor. Not only is the variation rather great between the point-conact and the junction devices and between germanium and silicon units, but it varies over a wide range between units of a given type by a given manufacturer. Thus, when the frequency range has been specified, serious consideration must be given to the type of transistor that is to be used. Further, it may be necessary for the designer to specify

characteristics of a given type of transistor which must fall within certain tolerances in order to assure the desired performance.

One criterion by which transistors may be compared as to frequency limitations is the alpha cut-off frequency. The alpha cut-off frequency is defined as the frequency at which the magnitude of alpha is reduced by 3 db from its low frequency value.

In general, point-contact transistors are useful to higher frequencies than are the junction types. The alpha cut-off frequency of the point-contact device may be rated as high as 50 megacycles against one or two megacycles for the usual germanium junction unit. The use of silicon in the junction transistor results in a somewhat increased upper limit, but this is still below what may be achieved with point-contact transistors. Very recently, Raytheon has announced the development of their CK-760, CK-761 and CK-762 for operation at 5, 10 and 20 megacycles. These are junction transistors.

To illustrate the deviation in alpha cut-off frequency within a group of transistors of a single type by a given manufacturer, the measurement of this parameter of 50 NPN, grown junction, germanium units revealed a spread from 160 kilocycles to 2.51 megacycles with most of the units lying between 600 kilocycles and 1.0 megacycle. The manufacturer listed the "average" as being 900 kilocycles.

A criterion more meaningful in the common-emitter configuration is the b cut-off frequency, which is defined as the frequency at which the magnitude of b is reduced by 3 db from its low frequency value. To a good approximation,

$$f_{bo} = f_{\alpha o}(1 - \alpha) \qquad (18)$$

From this equation, it is apparent that for transistors with alpha near unity, the b cut-off frequency may be very low.

The foregoing paragraphs indicate that the frequency may dictate the type transistor to be employed and further that it may even influence the configuration of the circuit. Fortunately, transistor oscillators will function with good results up to three or four times the alpha or b cut-off frequency in the appropriate configuration. Exactly how high above these frequencies satisfactory operation may be obtained depends to a fairly large extent upon how heavily the oscillator is loaded. Demanding greater power from the circuit will reduce its upper frequency limit.

At the upper limit of the frequency range, the phase shifts through the transistor itself, due primarily to alpha and the collector impedance becoming complex, become so great that oscillations are no longer sustained. Phase-correction networks or elements may sometimes be employed to improve the high frequency performance and raise the upper limit. One illustration of this is shown in Figure 23 where C_e

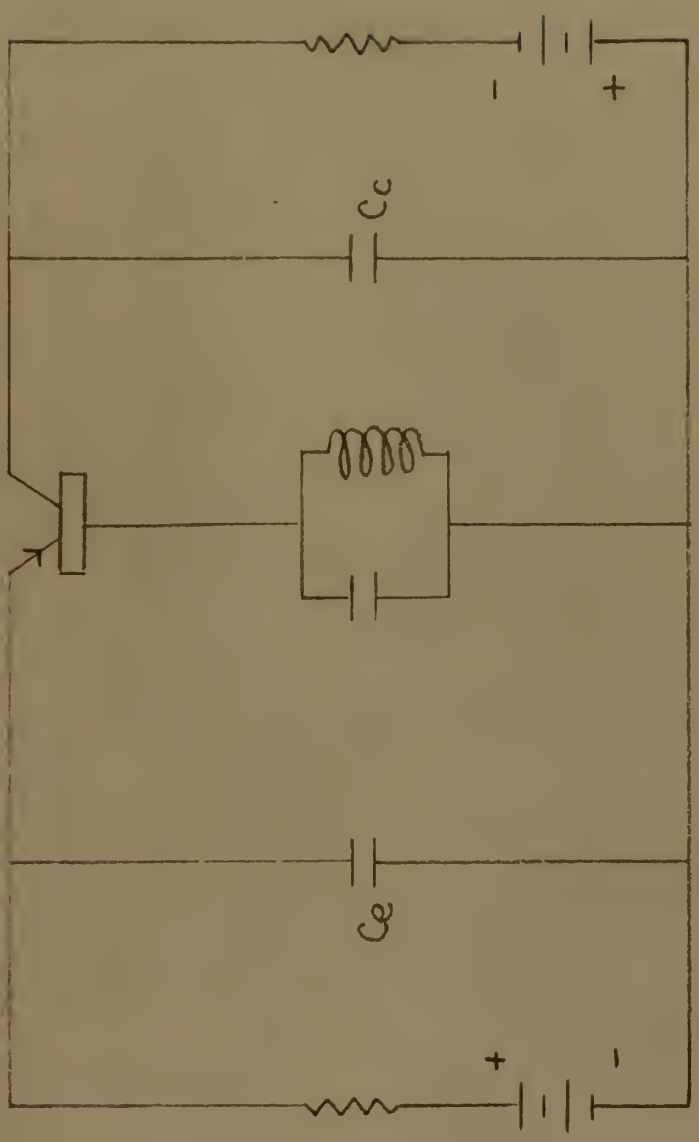


Figure 23

and C_c are employed to improve the operation of the negative-resistance oscillator from Figure 2. Optimum values exist for C_e and C_c [2].

3. Frequency Stability.

Transistor oscillators are subject to frequency variations from all of the sources encountered in tube circuits. However, the variations are considerably greater in the transistor circuit for a given condition. Germanium transistors are extremely temperature sensitive with the silicon device being somewhat more stable. In addition, variations in the biasing source will result in fairly large frequency deviations. In fact, one method of achieving frequency modulation consists of varying the collector or emitter bias.

Crystal-controlled operation improves the stability as it does with vacuum-tube circuits and with the crystal in a suitable oven, transistor crystal oscillators may be made to approach the stability of their vacuum-tube counterpart. Several crystal-controlled oscillators were shown in Chapter III and they will be further discussed in Chapter V.

In addition to using a crystal, the stability of the oscillator may be improved greatly by the use of a properly designed biasing network. This network serves to minimize the effects of variations in the collector saturation current of the transistor. This saturation current is a function of temperature and the collector potential and vari-

ations in it produce frequency variations. The two methods found most satisfactory in arriving at a proper network for this purpose will be covered in detail in the next chapter.

Llewellyn [9] did considerable work on this problem as it pertained to vacuum-tube oscillators of the Hartley and Colpitts types. Liben [2] extends this work to transistor oscillators in the frequency range where the transistor parameters may be considered real quantities.

Using the equivalent circuit of Figure 24, it may be shown that the frequency will be independent of variations in bias currents or of changes in a resistive load connected in series with Z_1 or Z_2 provided only that

$$(r_1 + r_b + r_e)(X_4 + X_m)^2 + (r_2 + r_e + r_c - r_m)(X_3 + X_m)^2 + (2r_e - r_m)(X_4 + X_m)(X_3 + X_m) = 0 \qquad (19)$$

where jX_m is the mutual impedance between X_3 and X_4.

With X_3 and X_4 properly adjusted, this condition is met and oscillations at a frequency independent of the bias current and load result.

4. Output Level.

Transistors, as vacuum tubes, dissipate power internally. This dissipation manifests itself as increased junction temperature. Since we know that temperature affects the transistor deleteriously, it is clear that some maximum value of dissipation exists which must not be exceeded.

Since the transistor is not a 100% efficient device,

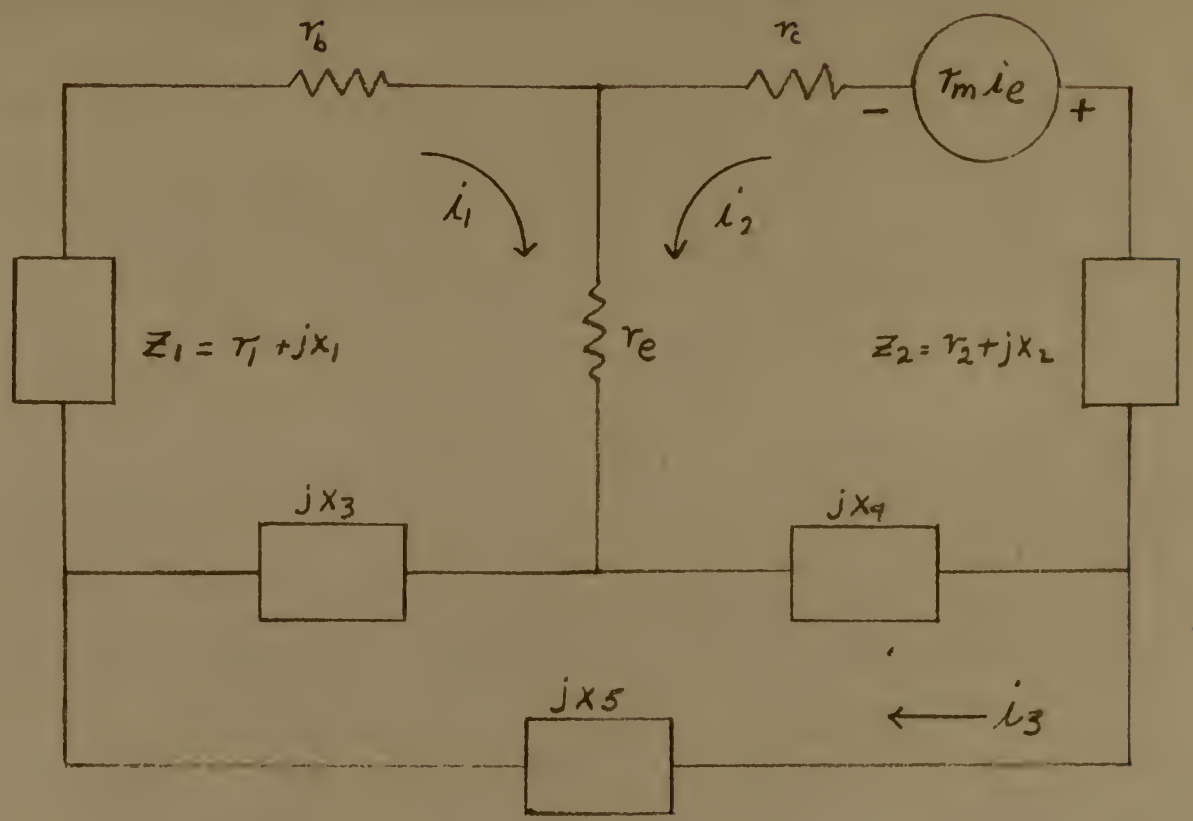


Figure 24

it follows that a given unit is capable of delivering only so much power. This power level will be a function of the ambient temperature, the circuit efficiency, the biasing point and the type transistor employed in the circuit.

Transistors capable of delivering several watts of power have been fabricated employing heat sinks or by circulating some coolant, air or a liquid. However, the more common transistor types are rated at only a fraction of a watt, usually in the neighborhood of 25 to 75 milliwatts for germanium units, and around 200 milliwatts for silicon.

These factors must be considered in selecting the transistor for the specific circuit under consideration.

The stability of the output level with variations in bias supply may be enhanced by a suitable bias stabilizing network and the proper choice of biasing point. The solution to the problem is discussed rather fully in Chapter V. Another approach to level stability, in the case where the output power level is not required to be near the maximum capability of the selected transistor, is the use of a thermistor across the output. As the output level tends to increase, the thermistor will exhibit a decreased resistance, loading the oscillator further and tending to reduce the output level. With decreasing output levels, the thermistor appears as an increased resistance, reducing the loading. This action represents a stabilizing influence and can be made to produce excellent results. It must be borne in mind

that the thermistor dissipates significant power and that this power must be furnished by the transistor without overloading. Caution must be exercized in the use of the thermistor in any environment in which temperature variations may be encountered since it is extremely sensitive in this regard.

In many circuit configurations, the replacement of the transistor will result in rather large changes in the output level. When this occurs, it is required that some means of varying the feedback be available if the circuit is not to be redesigned for each transistor. If some latitude in the level is permissible, it is possible that the specification of transistor parameters within given tolerances may permit interchanging transistors without any circuit changes. However, the selection of transistors with parameters falling within close tolerances is an expensive process since the tolerances set by a manufacturer are rather broad within a given type. This cost may be reduced if sufficient statistical data is available to permit the engineer to select the characteristics such that they include the maximum number of units. At the present state of the fabricating technology, however, this statistical data is limited and frequently represents the measurements of a relatively small number of transistors, and even these may be invalidated by some change in the manufacturing processes.

5. Impedance Matching.

In the preceding article, the output level was assumed to be the level of the power delivered to the load, implying a power oscillator. This is not necessarily the case and the discussion is valid however small the power delivered. However, assuming the oscillator to be driving a transistor amplifier or modulator or other power consuming device, it is necessary that its output impedance be matched for the maximum power output. This raises again the lack of uniformity among transistors of a giventype, and the effects of temperature. The collector resistance of the germanium transistors falls rather sharply with moderate increases in temperature, the silicon devices being somewhat more stable until much higher temperatures are reached. Since the output impedance is a function of the feedback ratio, in most cases it will be satisfactory to adjust the feedback to provide the required output level, accepting the attendant loss due to mismatch. However, it is possible that the output impedance of the oscillator may be a rather critical factor. For example, consider the case where a transistor amplifier is being driven by the oscillator. Any change in impedance seen by the amplifier input circuit will result in variations in the output impedance of the amplifier causing a further mismatch to occur at this point. When this output level must be fixed, an unacceptable situation may be encountered; not because the oscillator is

developing insufficient power but rather because of the loss of power due to the mismatch reflected across the amplifier. Again the solution lies in the proper selection of transistors for replacements and the design of the proper biasing network to stabilize for temperature variations. It should be noted that this situation may arise whether or not the oscillator output impedance is matched. In very critical circuits, it may be necessary to provide some degree of isolation between the two circuits if interchangeability of transistors is to be achieved without a great deal of circuit readjustment.

6. Noise.

Noise in transistors is of two types. One is similar to the familiar Johnson noise and the other a rather erratic noise [10]. As in the case of vacuum-tube circuits, the transistor circuit may be represented by a passive network with appropriate generators included. If the noise is to be taken into account, two noise generators must be included. The two noise sources are not entirely incoherent.

Noise in the transistor depends upon frequency, collector voltage, emitter current and the value of resistance seen by the input circuit of the transistor [11]. This noise is found to fall off at about three db per octave with increasing frequency. More precisely, it is proportional to frequency raised to the -1.1 power. It increases with an increase in either collector voltage or emitter current.

There is an optimum value of the resistance seen by the transistor input circuit. This value of resistance is not the value which will give the greatest gain in an amplifier circuit. The noise figures for the three standard configurations vary slightly with the optimum resistance in the input circuit. However, this variation is generally masked by the differences which arise from the necessity of using different resistance values to provide an acceptable gain in the different configurations. Using values of resistance in the input circuit that are 0.1 or 10.0 times the optimum value increases the noise figure by about five db.

Noise figures in transistors are generally taken at one kilocycle and with the biasing listed under "Typical Operating Characteristics" on the data sheets. Point-contact transistors appear to be the least desireable from the standpoint of noise with noise figures in the neighborhood of 50 to 70 db. Junction transistors are considerably better with noise figures of about 20 or 30 db. Shockley [12] states that there appears to be no physical principle which will prevent lowering the noise greatly in transistors. As evidence of the validity of this statement, Raytheon has developed a low noise PNP junction device with a noise figure of 12 db.

The specification of a low noise figure, then, may cause the designer to avoid the use of point-contact transistors in critical circuits, as for example in the first stages of

of a superheterodyne receiver.

7. Miscellaneous Considerations.

While the interchangeability problem has been discussed from several aspects, little has been mentioned comparing the point-contact and the junction or germanium and silicon devices. Experience indicates that for the purpose discussed here, junction transistors are somewhat more uniform than the point-contact. Also, the uniformity between silicon units seems to be better than for germanium.

Cost-wise, the germanium transistor is still considerably less expensive than is the silicon. Also, the junction device is a little lower in price than the point-contact.

CHAPTER V

DESIGN OF A TRANSISTOR OSCILLATOR

1. General.

In this chapter, an oscillator will be designed to fulfill a specific requirement. The circuit selected, the design procedure employed and the results obtained are presented as but one solution to a specific problem, and not necessarily the best. All characteristics have not been optimized, but this is frequently not necessary. In most practical engineering efforts, both time and component cost must be taken into consideration and weighed against the advantages of an improved final product obtained at the expense of more time and money. This is applicable in this instance, which was undertaken as a part of a commercial venture in a highly competitive field. It might be emphasized however, that the fact that the circuit may not be the optimum from a purely technical point of view does not imply that the design specifications are not met. This is not the case. Design specifications are usually written such as to insure some minimum standard of performance by the apparatus under consideration. Violation of the specifications then usually results in the degradation of the performance to an unacceptable extent.

2. Design Specifications.

The power required of the oscillator was nine milliwatts

into a load which was resistive and equivalent to 1000 ohms. This level was to vary not more than plus or minus three db over the temperature range 0° to 50° centigrade with 90% humidity. Initially, the specifications required operation from a 48-volt, unregulated, power supply with the "normal" voltage lying between 46 and 52 volts and the voltage for "emergency" operation extending from 42 to 54 volts. Later in the program, the nominal voltage was changed to 130 volts with the "normal" range from 125 to 136 volts and "emergency" 110 to 138 volts. The three db variation in level was not to be exceeded over the "normal" range but some degradation was permissible in the "emergency" region which extended beyond "normal".

In addition, it was specified that the Texas Instrument Corp. Type 200 transistor be used. This is an NPN, grown-junction transistor and its characteristics are shown in Appendix III. This choice followed from the decision that the junction transistor was superior to the point-contact for this application and from cost considerations, including contractual arrangements, which gave the TI-200 a relatively large cost advantage.

The frequency was specified as 52.000 kilocycles, plus or minus one cycle under the temperature and voltage conditions listed above. The oscillator was to be crystal controlled with an NT-cut quartz crystal operating in a 75° centigrade oven.

Harmonic content of only the second order was considered detrimental since other harmonics fell well beyond the cut-off frequency of the filters in the system. The second harmonic was to be as low as possible, not to exceed 20 db below the fundamental, to avoid increasing filter requirements and costs.

3. Circuit Selection.

The choice of a specific circuit was influence by certain considerations in addition to the specifications. Quantity production of the system was anticipated. If this is to be reasonably economical, standard components with standard tolerances are desireable. In addition, replacement of various components in the circuit due to failure should not be a critical operation. These considerations led to the choice of a circuit which would permit the crystal to operate in its series mode rather than between the series and parallel modes as is frequently the case. This series-mode operation tends to make the frequency much more independent of the remainder of the circuit.

The choice of configuration between common-base and common-emitter was made by designing both and determining experimentally which of the two offered the greater frequency and output level stability.

These two circuits are shown in Figures 25 and 26. Frequency stability was found to be equally good in the two circuits. However, the common-emitter circuit proved to be the

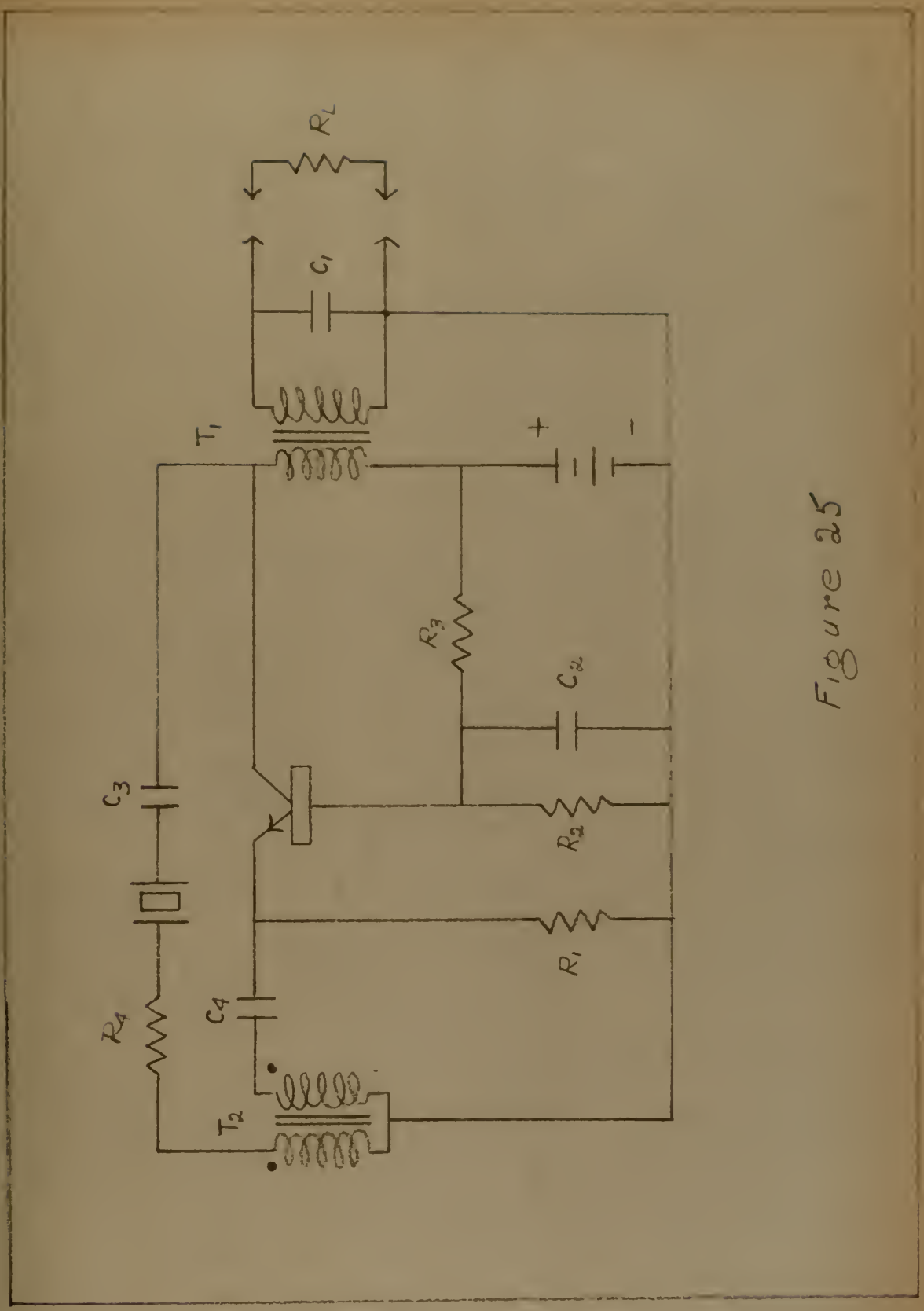


Figure 25

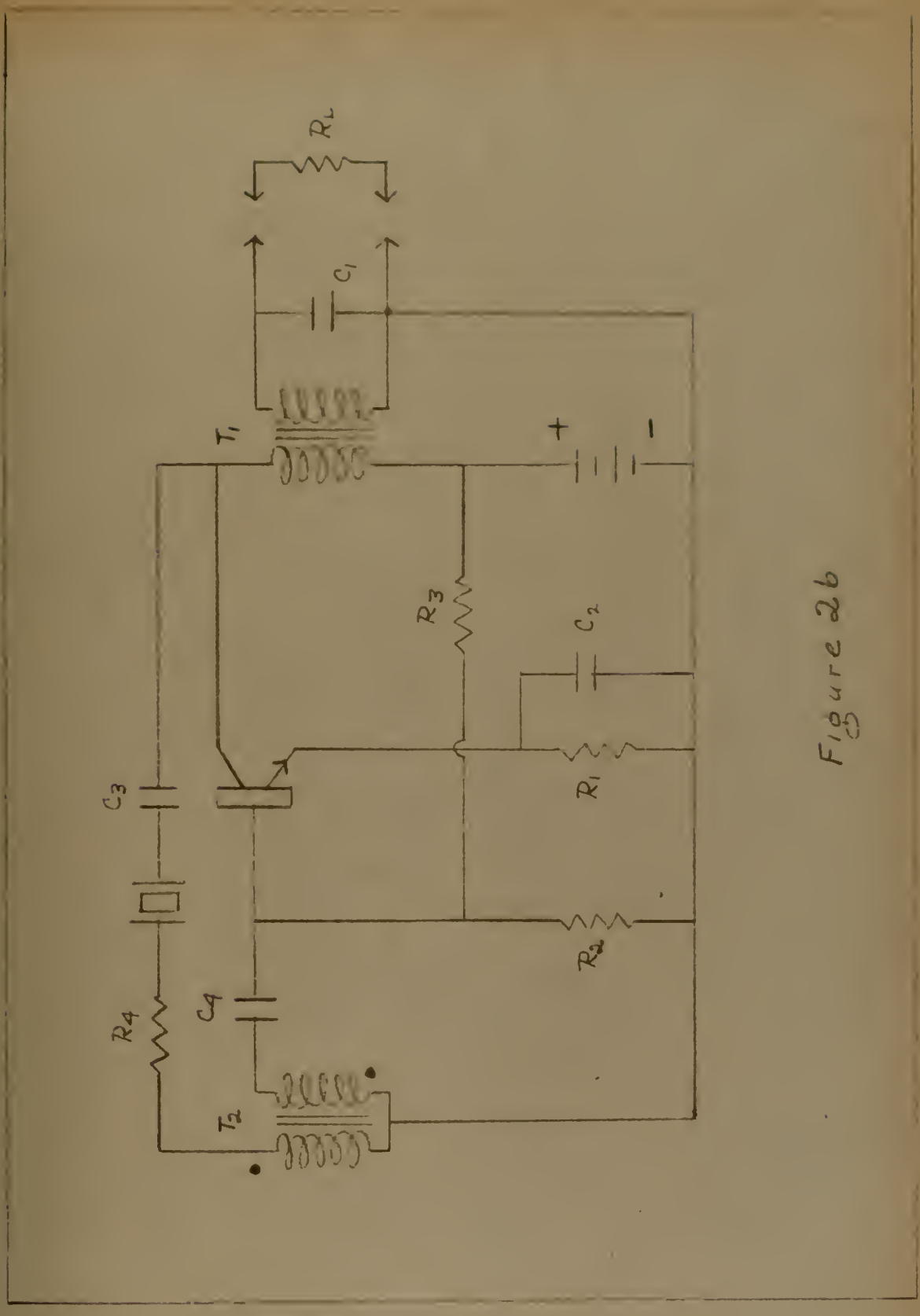


Figure 2b

more stable with temperature variations, as may be seen from Figures 27 and 29. The collector current variations, which will be discussed later in the chapter, are shown in Figures 28 and 30. These test were made for three different TI-200 transistors to obtain some idea of the output level variation that might be expected in both circuits with different transistors. At the same time, the data displayed in Figures 31 and 32 were collected to indicate performance of the common-emitter circuit with different transistors under power supply voltage variations and temperature changes. The results of these tests indicate the feasibility of this circuit for the purpose at hand.

Before settling upon the final circuit configuration, one other cost-reducing possibility was investigated. This was the possibility of combining transformers T_1 and T_2 in the circuit of Figure 26 and using a single transformer. This results in the circuit of Figure 33. Except for relatively minor variations, this circuit was adopted as final. These minor variations were required to improve the output level stability and to reduce the second harmonic content in the output and will be discussed fully later in the chapter. Component values have not been given thus far since we are interested here only in arriving at the appropriate circuit. These values will come to light in their logical sequence in the design procedure.

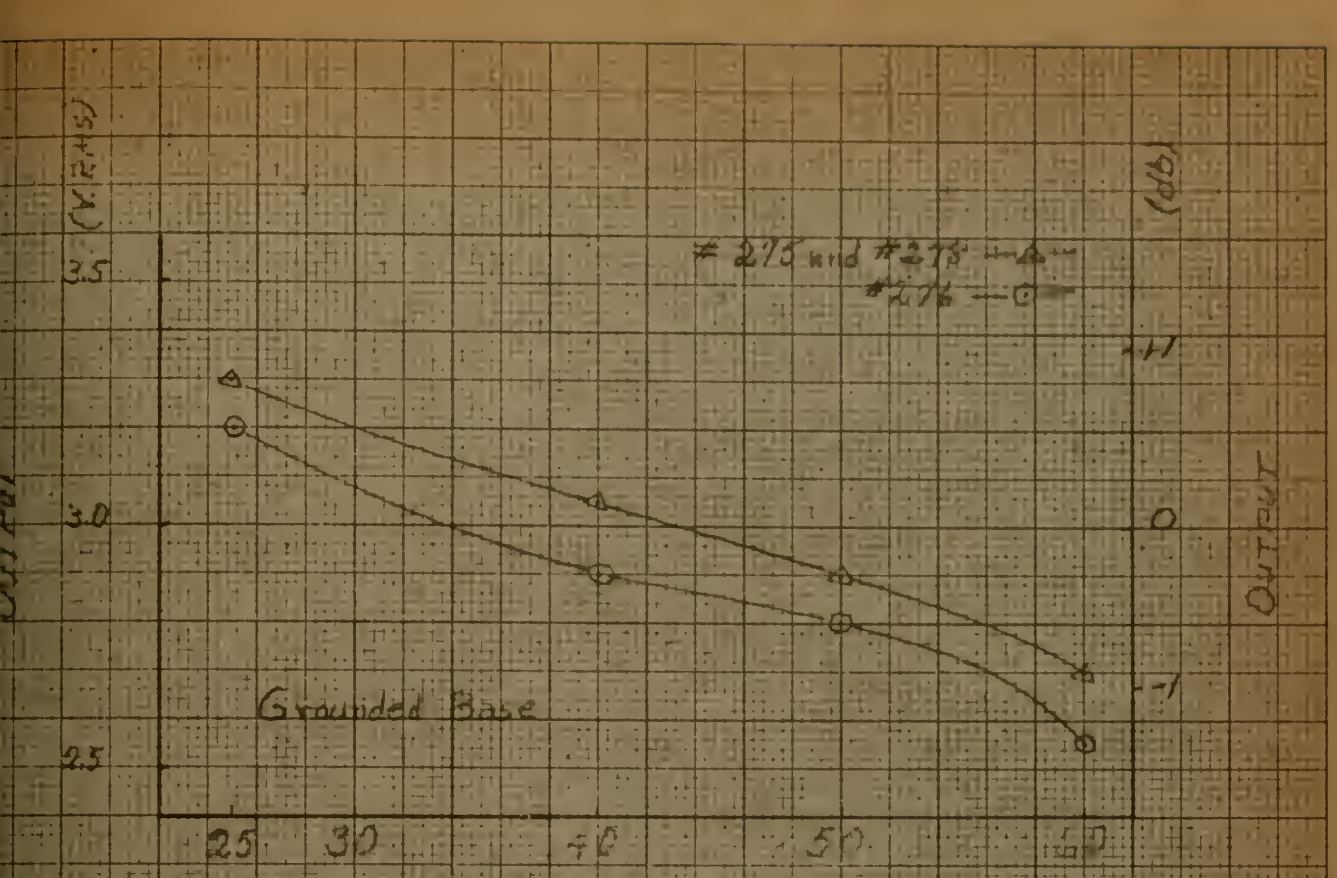


Figure 27

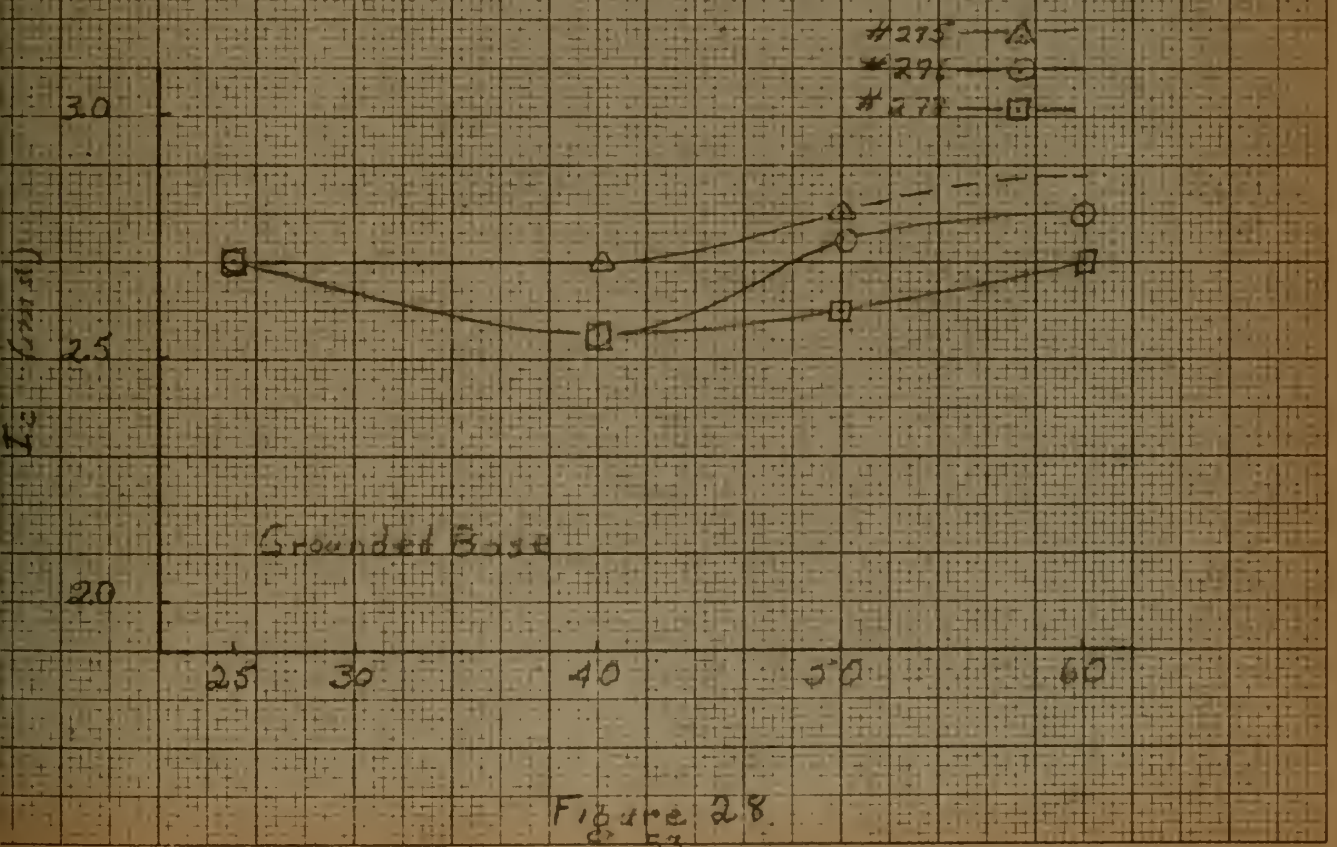


Figure 28.

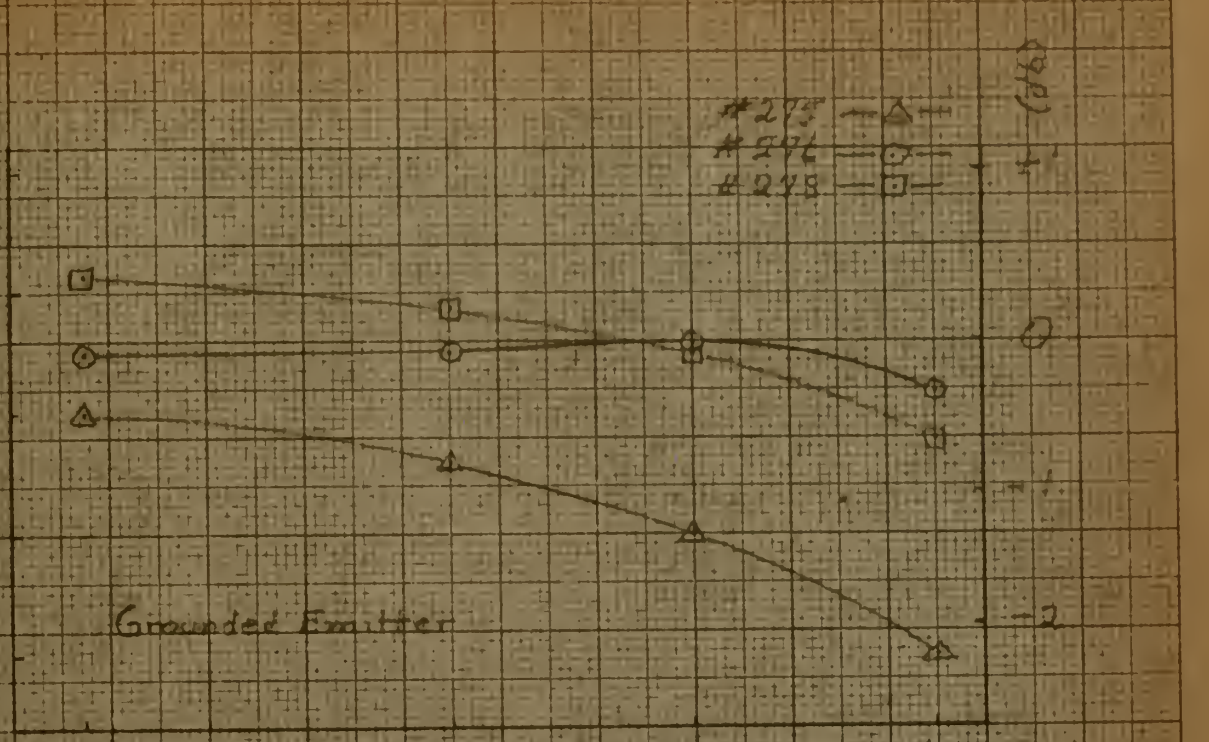


Figure 29

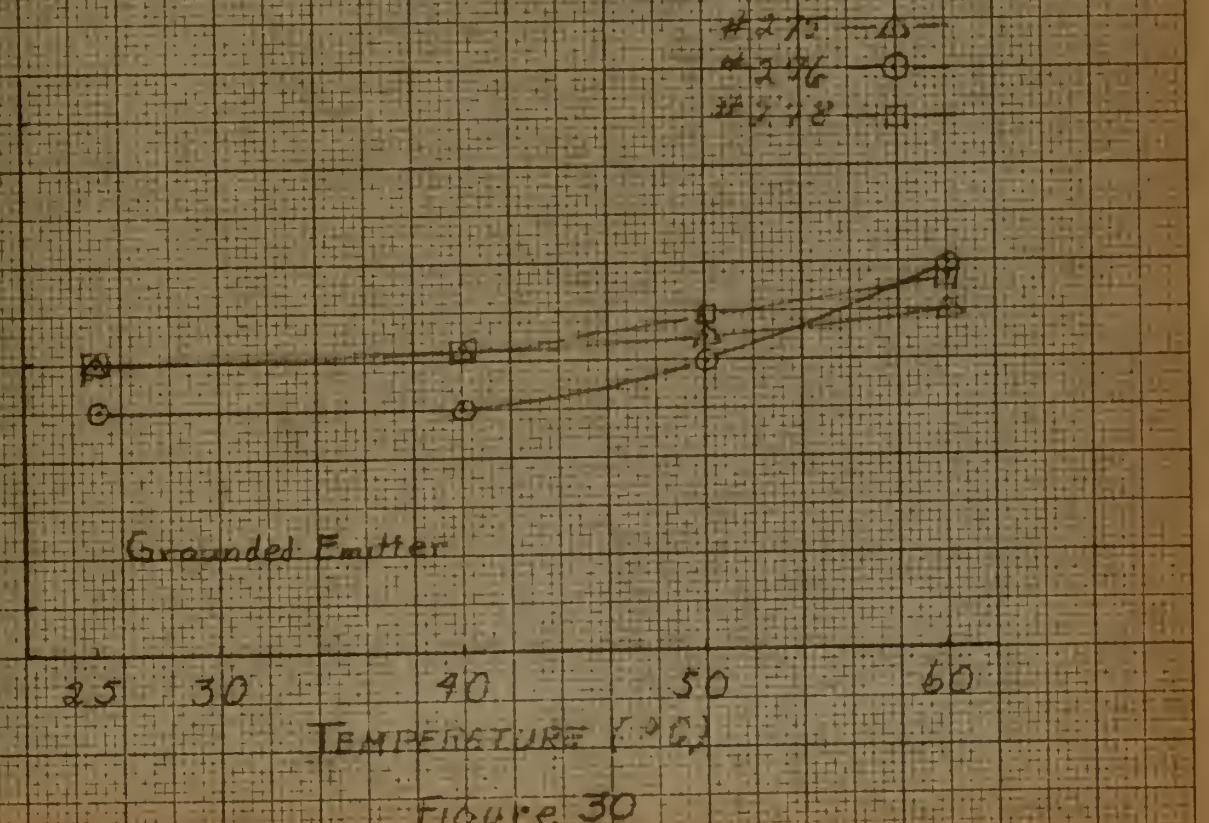


Figure 30

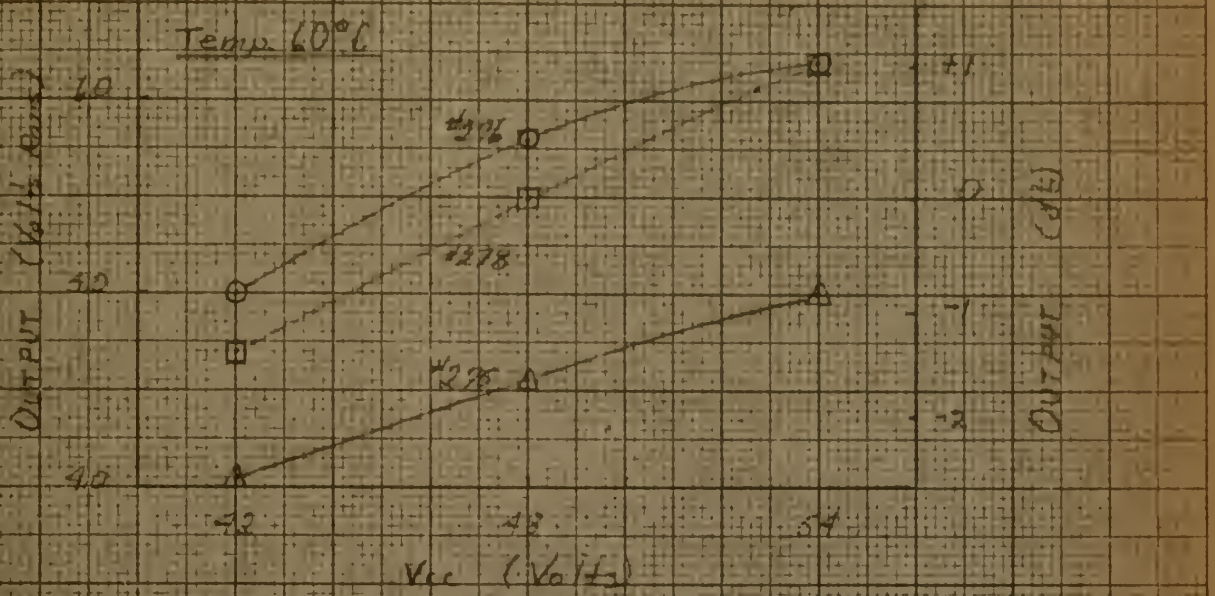


Figure 31

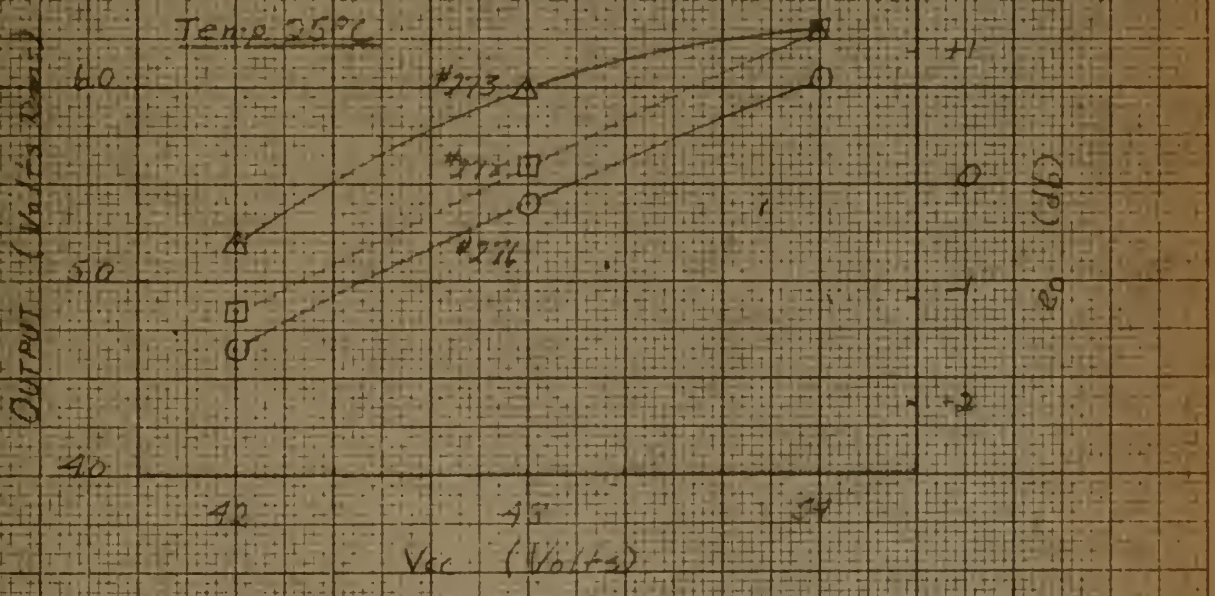


Figure 32

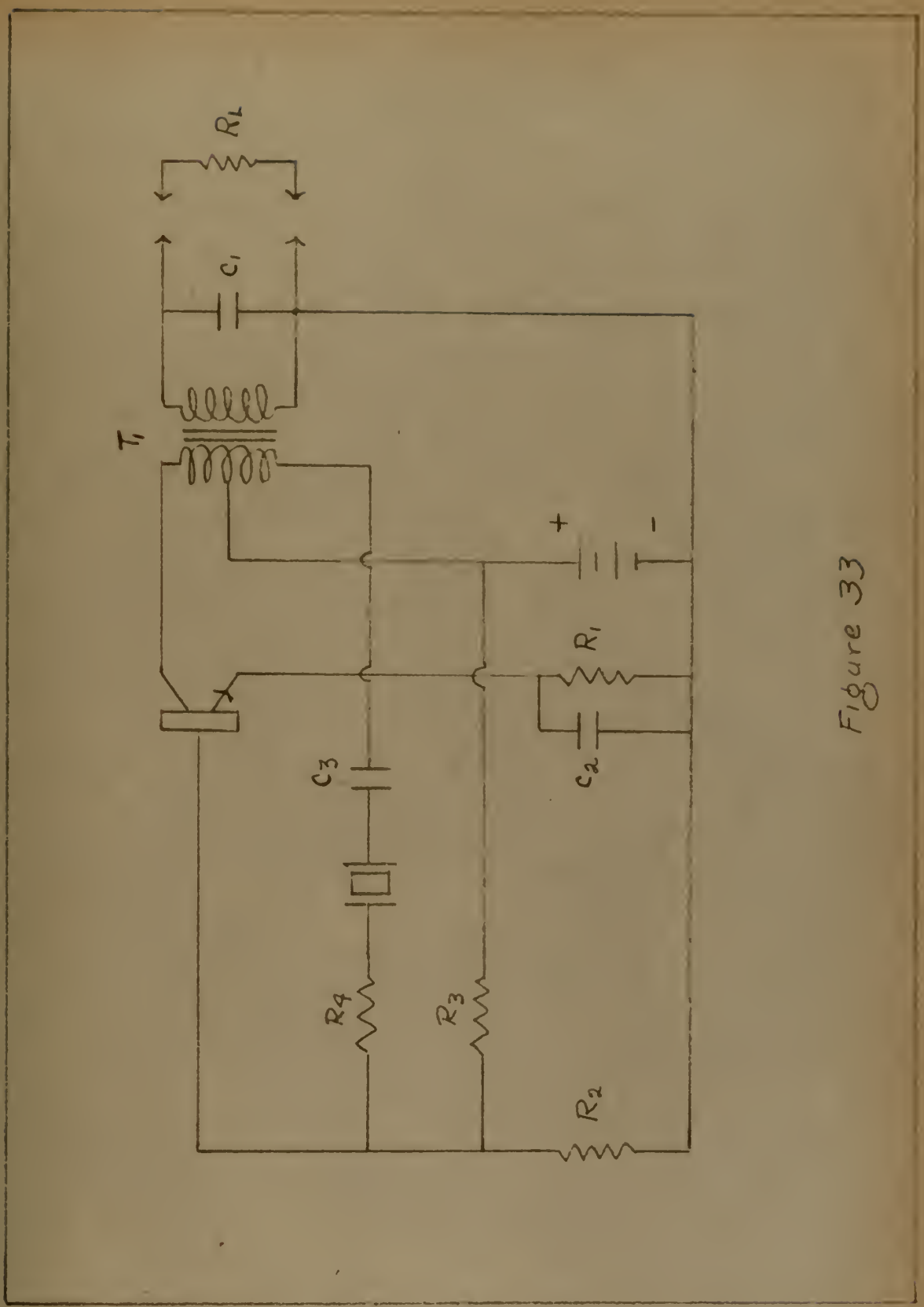


Figure 33

4. Design Procedure.

Since 52 kilocycles is in the vicinity of the b cut-off frequency for a large number of individual transistors, it is to be expected that the current gain of these units will be down three db from the low-frequency value. The power gain is down at this frequency to approximately one half the low-frequency gain. The low-frequency output characteristics supplied by the manufacturer and shown in Appendix III lose much of their accuracy under these conditions.

Since the maximum power output that may be obtained at the collector without exceeding the recommended maximum collector voltage of 30 volts or collector current of five milliamperes is 18.75 milliwatts, it was decided to design for this maximum which is about twice the required load power .

Some of the "excess" power will be required for feedback and some dissipated in the transformer and tuned circuit.

With this decided, the operating point was chosen with the collector voltage at 15 volts and the collector current at 2.5 milliamperes. A load line representing 6000 ohms drawn through this operating point intersects the output characteristic axes at 30 volts and 5 milliamperes. This will permit a current swing of 2.5 milliamperes peak, or about 1.77 ma. rms. For the transistor with an alpha of

0.90, b will be equal to nine at low frequencies. Near the cut-off frequency this is reduced by three db to 6.36. This represents the lowest b expected to be encountered and therefor will be used, following the reasoning that if sufficient feedback is provided for the lowest value of b, greater values will function satisfactorily.

Since the collector current is b times the base current it is necessary then to feed back a current equal to i_c/b to the base circuit. For the case at hand this will be 1.77/6.36 or 0.278 ma. rms. These values fix the impedance ratios of the transformer T_1. A 6000-ohm load is desired between the collector and the center-tap of the transformer. A standard transformer was available with an impedance ratio of 16,000 ohms to 600 ohms at this frequency. This provides a 6.67 K-ohm to one K-ohm impedance ratio between the upper half of the primary and the secondary and a 6.67 to 6.67 impedance ratio between the two halves of the primary.

With this transformer and an a-c. collector current of 1.77 ma., a simple calculation shows that the resistance of the feedback path should be 35.5 K-ohms to provide 0.28 ma. feedback current. This resistance and the load resistance of one K-ohm appear across the upper half of the transformer primary as 35.5 K-ohms in parallel with 6.67 K-ohms. This gives an equivalent resistance of 5.62 K-ohms which is relatively close to the desired six K-ohms load desired. A closer match would entail the design of a special trans-

former, which is undesireable.

The resistance of the feedback path is a reasonable value since the resistance of the crystal may be expected to be between 20 and 30 K-ohms. The capacitor is large and effectively a short circuit at this frequency. The resistor R_4 was made variable over a range of zero to 50 K-ohms to avoid excessive feedback when transistors with a current gain greater than nine were employed or when the b cut-off frequency was sufficiently high as to avoid the three db decrease in b. Since no variables were permitted in this circuit for the final configuration, R_4 must be selected in the production phase after the specific crystal and transistor have been placed in the circuit.

With 0.28 ma. feedback current, 1.77 - 0.28 or 1.49 ma. will be transformed to the secondary by the square-root of the impedance ratio. Thus, there appears in the secondary $\sqrt{6.67}$ x 1.49 or 3.58 ma. If this were all flowing through the load, 14.8 milliwatts would be delivered. However, circuit losses will account for some of this as was pointed out earlier.

The capacitor C_1 was chosen to tune the primary of the transformer to resonance. This was done experimentally and the value was found to be 300 micromicrofarads. The reason for placing the capacitor in the secondary rather than the primary was the large value of inductance of the primary. This reduced the value of capacitance required for resonance

to the point where the stray capacitances of the transformer and the remainder of the circuit were a significant part of the tuning capacitance. In placing it in the secondary the value of C_1 was increased by the impedance ratio.

The resistors R_1, R_2 and R_3 provide the biasing network and set the operating point. The capacitor C_2 is simply a by-pass capacitor and 0.1 microfarad was sufficiently large.

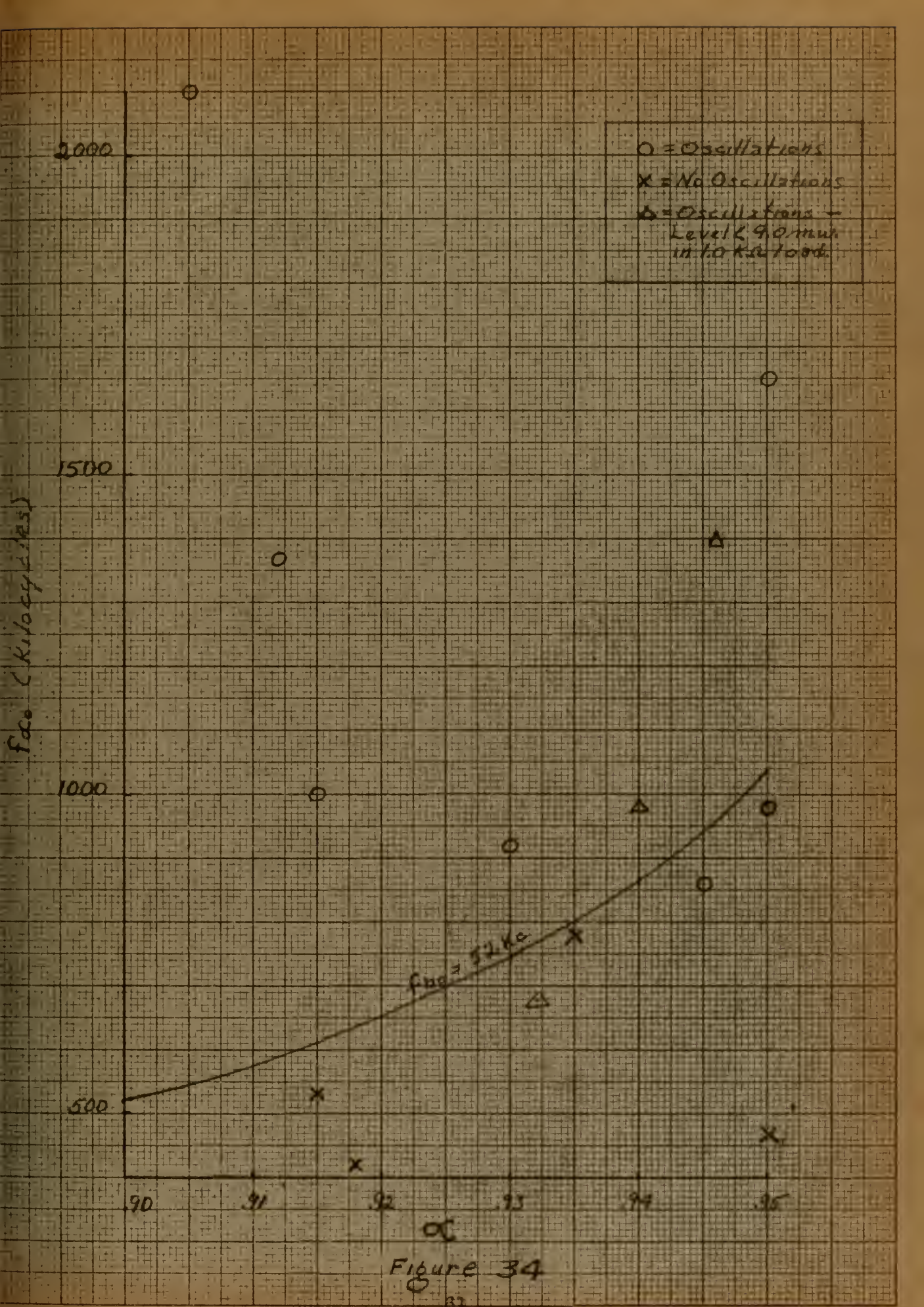
The biasing network was computed following the method of Shea [3] utilizing a stabilizing factor of two and a power supply voltage of 130 volts. An average alpha was chosen as 0.93 for these computations. This gave the values listed below.

Resistor	Computed Value	Nearest RMA Value
R_1	42.75 K-ohms	43.0 K-ohms
R_2	1.113 Megohms	1.10 Megohms
R_3	52.0 K-ohms	51.0 K-ohms

The RMA values were used in the circuit and deposited-carbon, high-stability, one-percent resistors specified.

5. Circuit Evaluation.

At this juncture, several transistors were placed in the circuit, one at a time, and their operation observed. As may be seen in Figure 34, 50% of them operated satisfactorily. All of those which failed to oscillate at all had a b cut-off frequency below 52 kilocycles per second



O = Oscillations
X = No Oscillations
Δ = Oscillations – Level < 9.0 mw. in 1.0 kΩ load.
2000
1500
1000
500
f_{α_0} (Kilocycles)
f_{α_0} = 52 Kc
.90
.91
.92
.93
.94
.95
α

Figure 34

which would reduce their current gain by more than the three db considered in the design.

Sufficient statistical data was not available to determine where the mean alpha of a large number of units would fall and thereby permit the specifying of a critical b cut-off frequency below which transistors would be rejected. It was decided to specify an alpha cut-off frequency of 900 kilocycles with alpha ranging from 0.90 to 0.95 pending the availability of further data.

Next, measurement of harmonic content was undertaken. The feedback resistor, R_4, was adjusted so that nine milliwatts of power were delivered to the load for each transistor. The results below indicate the level of the second harmonic with reference to the fundamental, in db.

Transistor #	233	235	243	245	247	228
Harmonic Level	-15	-15	-15	-14	-14	-13

This level is unsatisfactory. To improve this situation, it was decided to reduce the L/C ratio of the tuned circuit in the output, thus securing better filtering at the expense of output level. This resulted in the circuit arrangement shown in Figure 35. The harmonic content is reduced with decreasing values of L_1. However, the output level becomes unsatisfactory as L_1 is decreased much below 1.0 millihenry. A value of 1.1 mh. was chosen and tuned with C_1 having a value of 0.008 microfarad.

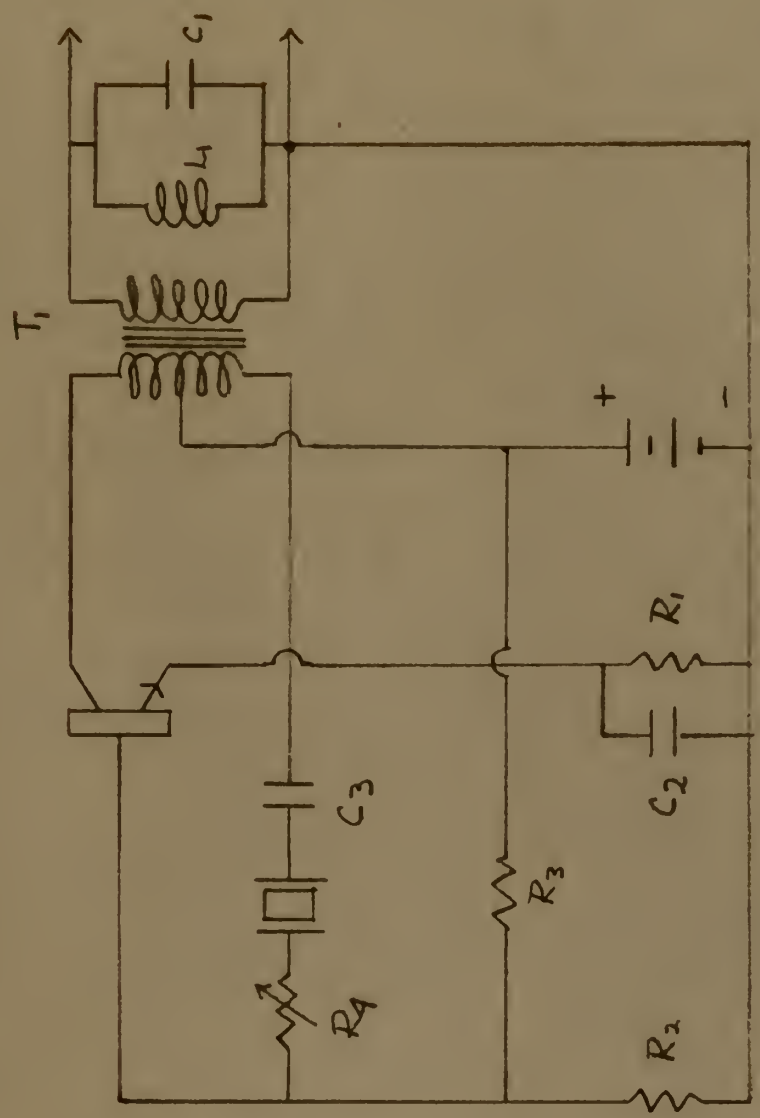


Figure 35

The harmonic measurements were again made with the feedback resistor adjusted for each transistor as in the previous test. Obvious improvement is noted.

Transistor #	233	235	243	245	247	228
Harmonic Level	-25.5	-25.0	-30.8	-20.5	-13.6	-27.3

Unit #247 was unable to produce a satisfactory level with this circuit but the improvement in harmonic content obtained obviated more stringent filter requirements and the additional cost of selection of transistors appears justified. This transistor is shown in Figure 34 as not providing sufficient output. It had an alpha of 0.943 and and alpha cut-off frequency of 1.4 mcs. From Figure 34, it should function properly and since it did not, it was assumed to have some other shortcoming.

With the harmonics now at a satisfactory level, tests were made to determine the performance of this circuit under conditions of varying temperature and supply voltage. The data obtained from these tests are tabulated in Table I. Zero db is arbitrarily taken as the level with 130 volts and 25° centigrade.

The results are generally unsatisfactory, with but a single transistor, #243, approaching the specifications. The reason for this is not obvious after the results depicted in Figures 31 and 32. It will be remembered, however, that the earlier circuit was stabilized for a supply

	25°C					40°C					50°C					60°C					Temp
TI-200	110	125	130	136	138	110	125	130	136	138	110	125	130	136	138	110	125	130	136	138	E
228	-1.4	-0.5	0	0.2	0.4	-1.6	-0.5	-0.1	0.3	0.5	-6.0	-5.0	-5.0	-5.0	-5.2	—	—	—	—	—	Output Level (db)
233	-1.4	-0.4	0	0.2	0.4	-1.8	-1.7	-1.4	-1.0	-0.8	-9.4	-3.3	-2.9	-2.6	-2.5	-7.1	-5.9	-5.7	-5.5	-5.3	
235	-1.1	-0.2	0	0.4	0.6	-0.8	0.2	0.6	0.9	1.1	-1.9	-0.9	-0.6	-0.5	-0.4	-10.8	-9.2	-8.8	-8.3	-7.9	
241	-1.5	-0.3	0	0.4	0.6	-2.1	-1.3	-1.1	-1.0	-0.8	-4.6	-3.7	-3.6	-3.7	-3.8	-11.4	-11.6	-11.6	-11.5	-11.8	
242	-1.7	-0.5	0	0.3	0.5	-1.8	-0.8	-1.1	-0.2	0.1	-3.5	-2.3	-1.9	-1.7	-1.5	-6.3	-5.6	-5.3	-5.0	-4.9	
243	-1.2	-0.3	0	0.4	0.6	-1.2	0.2	0.7	0.9	1.0	-2.0	-0.5	-0.2	0	0.2	-4.0	-3.4	-2.2	-1.7	-1.5	
245	-1.1	-0.2	0	0.4	0.5	-1.3	0.1	0.6	0.8	0.8	-4.5	-4.5	-4.5	-4.7	-4.9	-13.1	-11.9	-11.5	-11.3	-11.1	

Table I

source of 48 volts and the present circuit for a supply voltage of 130. While the values of the components were not listed for the first circuit, R_2, the resistor between the base and ground had a value of about 50 K-ohms. The circuit of Figure 35 has R_2 equal to 1.1 megohms. Remembering that the base current flowing through this resistor is the difference between the collector and emitter currents, it is seen that any variable which changes the collector current in a way different from the manner in which the emitter current varies will cause a change in the base current. Both temperature and voltage variations result in base current changes. Through the larger resistor, these variations, though small, become of greater significance.

With this in mind, a second method of bias stabilization was employed. Here, the method of Sherr and Kwap [13] was extended and appears as Appendix IV. The resulting circuit is shown in Figure 36. An additional resistor and capacitor are added to the circuit. The details of the calculations are included in Appendix IV with the results shown below.

Resistor	Computed Value	Nearest RMA Value
R_1	42.5 K-ohms	43.0 K-ohms
R_2	131.5 K-ohms	130.0 K-ohms
R_3	48.1 K-ohms	47.0 K-ohms
R_4	4.7 K-ohms	4.7 K-ohms

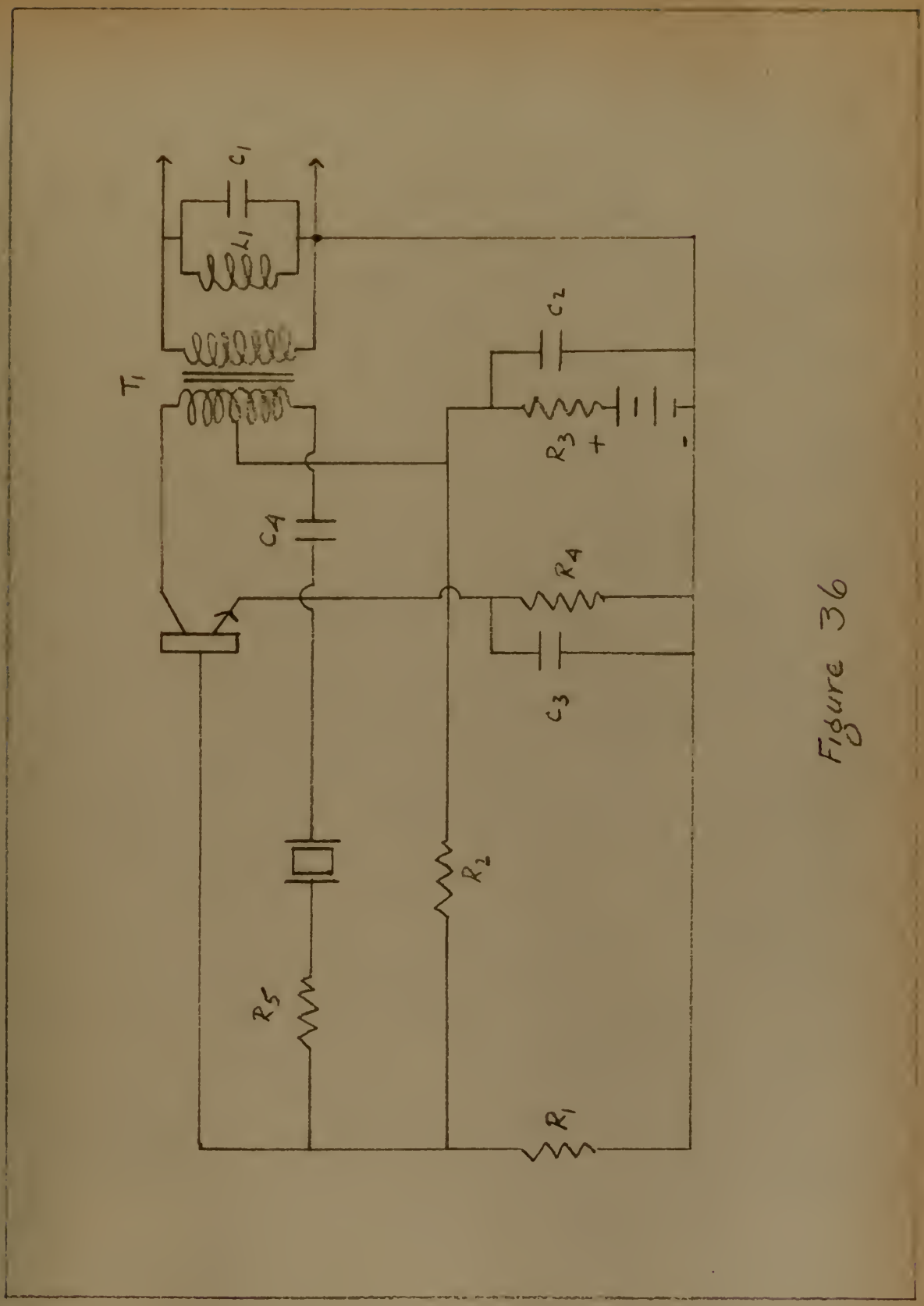


Figure 36

Capacitors C_2 and C_3 are both by-pass capacitors and have a value of 0.1 microfarad.

This method has the advantage of not only reducing greatly the value of R_1 but of adding R_3. Variations in the collector current which result in the voltage drop across R_1 being changed now will also produce a drop across R_3 which is compensatory.

The temperature and voltage variation tests were made again under the same conditions as before. Table II shows the results. A comparison of Tables I and II indicates the superiority of this second stabilizing method.

In the configuration of Figure 36, the circuit was accepted for the construction of the prototype models.

While the variation between 25° and 50° centigrade is greater than the three db variation in some instances, it should be recalled that 25° was arbitrarily taken as the reference and a temperature of 35° is much more likely to be the mean temperature and this brings the performance into agreement with the specifications for temperature increases. The necessary equipment was not available at the time to make measurements down to 0° centigrade but no difficulty is to be anticipated at this end. Humidity tests were not performed since results of these tests also reflect the adequacy of the packaging of the circuit and are made after the prototypes are available.

Frequency measurements were made continuously through-

	25°C					40°C					50°C					60°C					Temp
TI-200	110	125	130	136	138	110	125	130	136	138	110	125	130	136	138	110	125	130	136	138	E
228	-1.3	-0.3	0	0.5	0.8	-0.5	0.8	1.3	1.8	2.0	0.6	1.7	3.2	3.6	3.8	1.1	2.2	2.5	3.0	3.2	Output Level (db)
833	-1.7	-0.5	0	0.2	0.6	-0.7	-0.1	0.1	0.6	0.8	-1.0	0.1	0.5	1.0	1.1	-1.1	0.2	0.5	1.0	1.2	
235	-1.5	-0.4	0	0.5	0.8	0	1.1	1.6	1.9	2.1	1.1	3.3	3.7	4.0	4.2	1.3	2.6	2.9	3.2	3.5	
241	-1.7	-0.4	0	0.5	0.8	-1.0	0.3	0.7	1.1	1.3	-0.5	0.8	1.2	1.6	1.8	1.2	1.4	1.7	1.9	2.2	
242	-1.7	-0.3	0	0.4	0.7	-0.6	0.4	0.8	1.2	1.5	-0.2	1.9	2.2	2.7	2.9	0.3	1.4	1.8	2.1	3.2	
243	-1.2	-0.4	0	0.6	0.9	0.2	1.3	1.7	2.1	2.4	1.1	2.3	2.6	3.1	3.3	1.3	2.5	2.8	3.1	3.5	
245	-1.8	-0.6	0	0.5	0.8	-2.9	-1.4	-0.7	-0.4	-0.2	-4.3	-2.3	-2.2	-1.7	-1.5	-3.8	-2.4	-2.2	-1.8	-1.7	

Table II

out the evaluation phase and no indication of frequency shift was observed under any of the conditions. Frequency measurements were made with a frequently calibrated Eput meter, a direct-reading events-counting instrument, which was accurate to within one half cycle.

6. Discussion.

The design technique employed involved both the theoretical and experimental approaches. As a first approximation, the theoretical method gives reasonably good results. However, certain aspects of the design require the use of experimental techniques. This should not be too surprising since this is generally the case even in vacuum-tube circuits. The degree to which the analytical approach is valid may seem somewhat surprising in the light of the wide range of characteristics of the transistors employed. The situation is entirely analogous to requiring a vacuum-tube circuit to perform satisfactorily under widely varying conditions with tubes of several different types placed in the circuit.

While this circuit is of limited interest in itself, it is felt that the general design technique illustrated may be applied to meet other requirements. Further, this example points up some of the problems encountered in designing a circuit for quantity production and gives one solution to them.

This approach was applied to a 68-kc and a 32-kc crystal oscillator with similar results.

CHAPTER VI

CONCLUSIONS

1. The Present State of the Art.

It is felt that the information presented in Chapter V justifies the conclusion that, even with present-day transistors, it is possible to design transistor circuits for quantity production. Admittedly, a considerable part of the design work is accomplished at the work bench rather than on the drawing board. Much of this is required because of the disparity among individual transistors of a specific type. In addition, the inherent inability of germanium transistors to function properly above 70 to 80 degrees centigrade limits the usefulness of this type unit. While silicon transistors may be made to operate satisfactorily up to around 150 degrees centigrade, their current high cost makes their use prohibitively expensive for many applications.

Junction transistors, with their relatively high junction capacitances are limited in frequency range to applications lying in and below the medium-frequency band. Point-contact transistors have been successfully employed in the high-frequency band.

At present, the noise figure of the transistor and the power handling capability compare rather poorly with the vacuum tube.

2. Future Developments.

Many of these shortcomings are in the process of being eliminated at the present time. The technology of fabrication is rapidly closing the tolerances of a given type of transistor. Also, the manufacturing processes are reducing the cost of silicon transistors. Eventually, it is to be expected that this type should cost no more than the germanium variety since silicon occurs infinitely more abundently in nature than does germanium.

With the basic transistor sufficiently well developed, attention is now being focused upon its refinement. Units with good noise figures and low junction capacitances have been made in experimental lots and much improvement may be expected along these lines. Also units rated at up to 40 watts have been produced, raising the power level possible from the transistor by several orders of magnitude. It is not to be expected that units comparable to large transmitting tubes will be developed, however.

The discussion thus far has considered only the point-contact and the junction transistors of the "triode" type. Other types have been developed beyond the experimental stage. Some of these, such as the PNPN, the PNIP, the double-base tetrode and the field-effect transistor, show much promise toward expanding the field of application of the device.

The transistor may be expected to replace the vacuum

tube in many of its applications. However, the conclusion should not be drawn that the tube is in danger of becoming obsolete. The role of the transistor will be to complement the vacuum tube rather than to replace it entirely.

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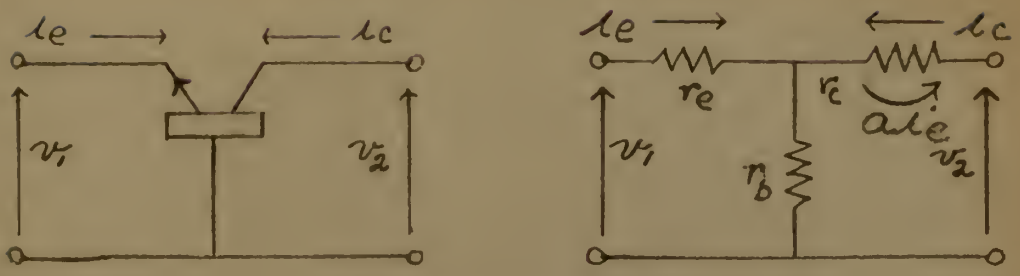
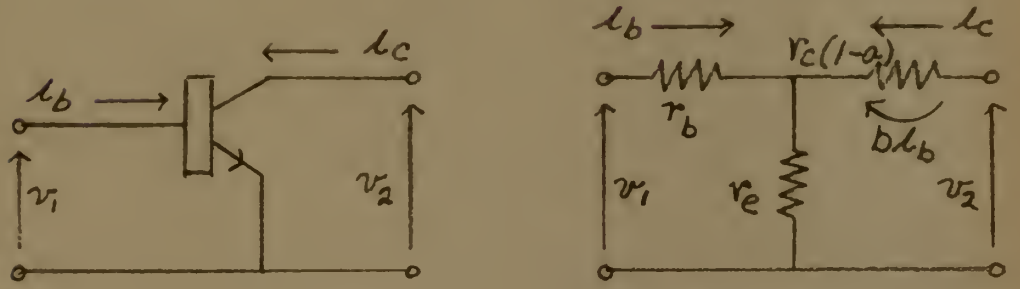
APPENDIX I

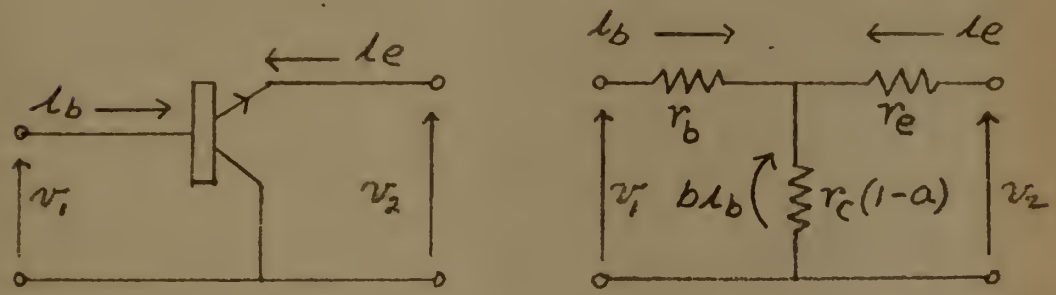
CURRENT GENERATOR EQUIVALENT "T" CIRCUITS



Common Base



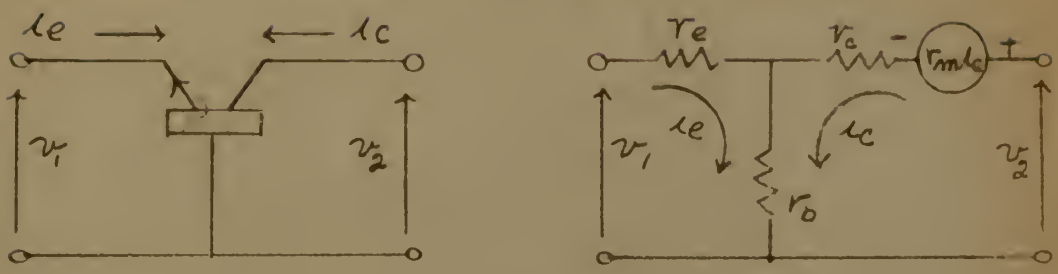
Common Emitter



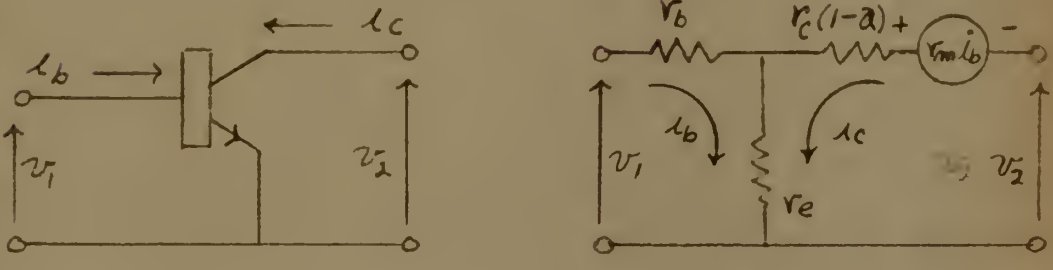
Common Collector

APPENDIX I

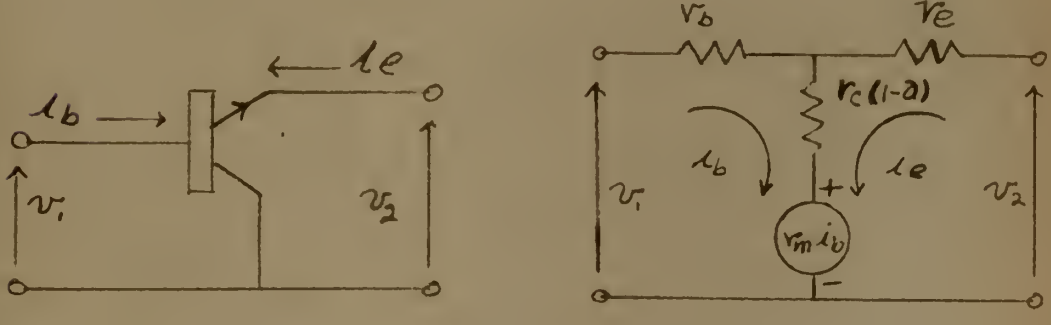
VOLTAGE GENERATOR EQUIVALENT "Z" CIRCUITS



Common Base



Common Emitter



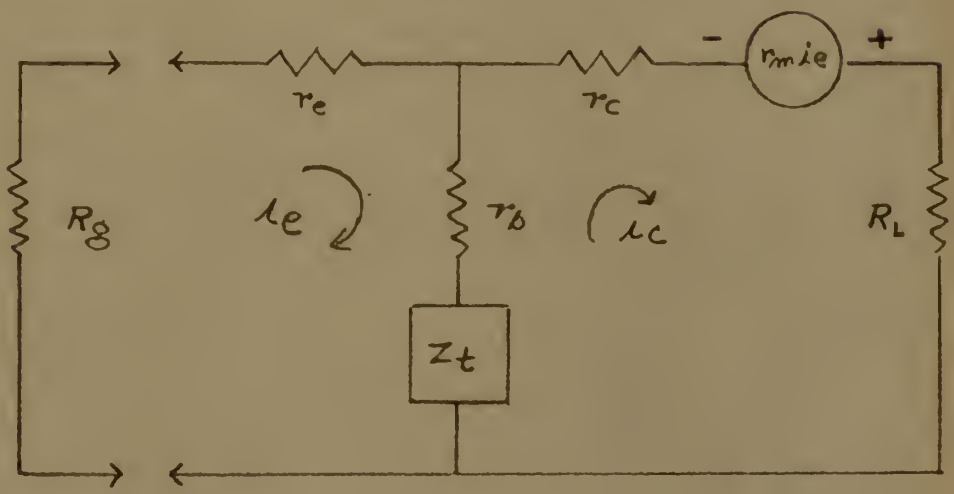
Common Collector

APPENDIX II

Derivation of Conditions for Negative Input or Output Resistance

Common Base Case

1. Negative Input Resistance



Writing the mesh equations for the circuit facing R_{g_0},

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = \begin{bmatrix} (r_e + r_b + Z_t) & -(r_b + Z_t) \\ -(r_b + Z_t + r_m) & (r_b + Z_t + r_c + R_L) \end{bmatrix} \begin{bmatrix} i_e \\ i_c \end{bmatrix}$$

$$Z_{in} = \frac{\Delta}{\Delta_{11}}$$

For oscillations to occur,

$$R_{g_0} < -Z_{in}$$

$$R_g < - \frac{r_e(z_t + r_b + r_c + R_L) + (z_t + r_b)(r_c + R_L) - r_m(z_t + r_b)}{z_t + r_b + r_c + R_L}$$

$$(R_g + r_e)(z_t + r_b + r_c + R_L) < - (z_t + r_b)(r_c + R_L - r_m)$$

$$z_t + r_b > \frac{(R_g + r_e)(r_c + R_L)}{r_m - (R_g + r_e + r_c + R_L)} \qquad (1)$$

This last expression is inequality (1) of Chapter II.

In the event Z_{in} is complex or R_g is replaced by a more general Z_g, the requirement for oscillation is modified to

$$Real[Z_g] < - Real[Z_{in}]$$

and the derivation carried out accordingly.

2. Negative Output Resistance.

The mesh equations for the circuit facing R_L are.

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = \begin{bmatrix} (R_g + r_e + r_b + z_t) & -(r_b + z_t) \\ -(r_b + z_t + r_m) & (z_t + r_b + r_c) \end{bmatrix} \begin{bmatrix} i_e \\ i_c \end{bmatrix}$$

$$Z_{out} = \frac{\Delta}{\Delta_{22}}$$

For oscillations

$$R_L < - Z_{out}$$

$$R_L < -\frac{(Z_t + r_b)(r_c + R_g + r_e - r_m) + r_c(R_g + r_e)}{Z_t + r_b + r_e + R_g}$$

$$(R_L + r_c)(R_g + r_e) < -(Z_t + r_b)(r_c + R_g + r_e + R_L - r_m)$$

$$Z_t + r_b > \frac{(R_g + r_e)(r_c + R_L)}{r_m - (R_g + r_e + r_c + R_L)} \qquad (2)$$

Again the result is inequality (1) of Chapter II, verifying the statement that making Z_t sufficiently large may result in either R_{in} or R_{out} becoming negative depending upon the values of R_L and R_g.

Similar derivations may be made for the common-emitter and common-collector circuits from their equivalent circuits.

APPENDIX III

TENTATIVE SPECIFICATIONS

for

Texas Instruments, Inc.

Type 200

Germanium Hermetically Sealed Junction Transistor

Electrical Data:

Ratings, Recommended Maximum:

Collector Voltage	30 volts
Collector Current	5 ma.
Collector Dissipation (25°C)	50 mw.
Ambient Temperature	50° C.

Average Characteristics (25°C):

Collector Voltage	5 volts
Emitter Current	-1 ma.
Collector Resistance (minimum)	0.4 megohms
Base Resistance	150 ohms
Emitter Resistance	22 ohms
Current Amplification Factor (Emitter Grounded) (minimum)	9
Collector Cut-off Current (maximum)	10 ua.
Collector Capacitance	15 uuf.
Noise[1] (V_c=2.5 v., I_c=-0.5 ma.)	26 db.
Frequency Cut-off[1]	900 kcs.

Typical Operating Characteristics[2] (25°C):

	Grounded Emitter	Grounded Base	Grounded Collector
Source Impedance	500 ohms	60 ohms	15-K ohms
Input Impedance	480 ohms	35 ohms	9.5 K-ohms
Load Impedance	20 K-ohms	100 K-ohms	500 ohms
Power Gain	37 db	30 db	12 db
Power Output[3]	2.15 mw	- - -	- - -
Distortion	6 %	- - -	- - -

Footnotes:

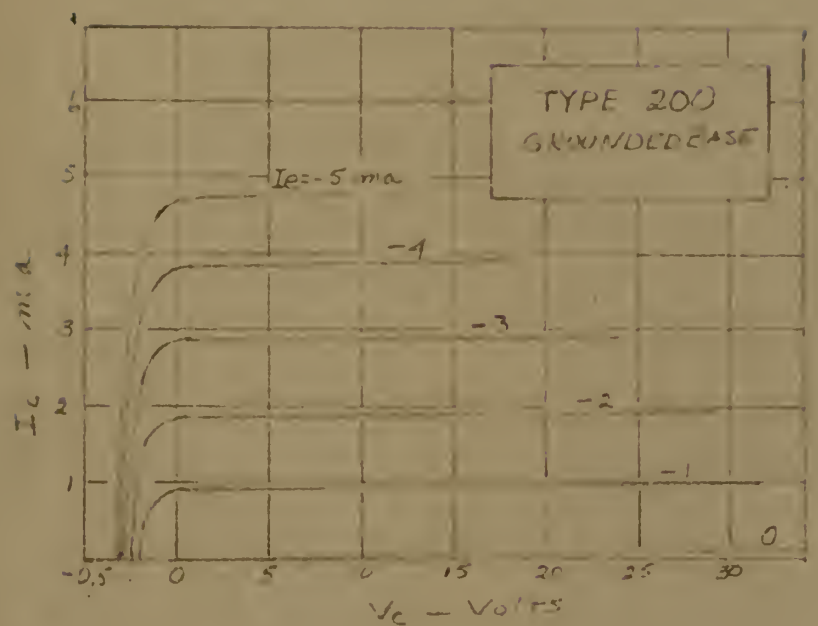
1. Noise and frequency cut-off are average and individual units may vary.

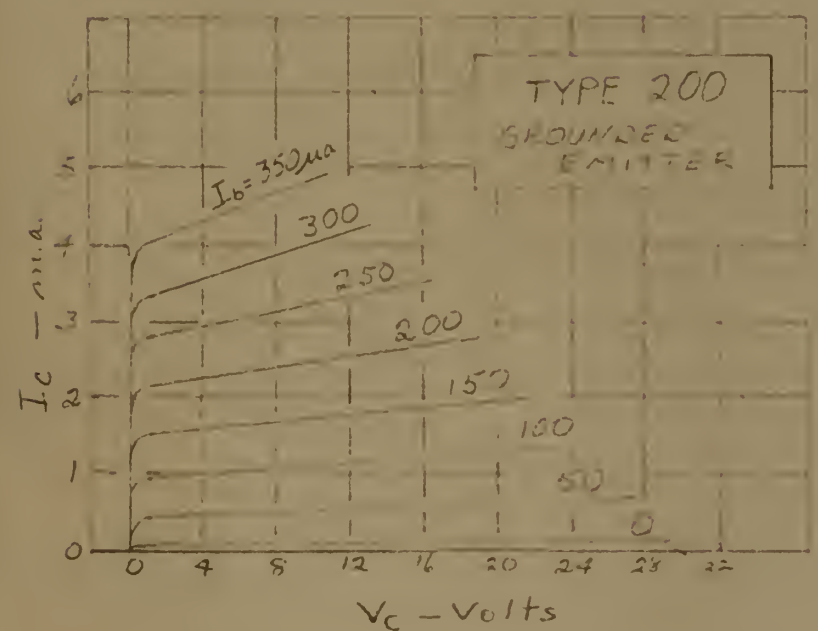
2. At test frequency of one kilocycle per second and V_c = 5 volts, I_e = -1.0 ma.

3. With a load impedance of 5000 ohms and high source impedance.

APPENDIX III

OUTPUT CHARACTERISTIC CURVES





APPENDIX IV

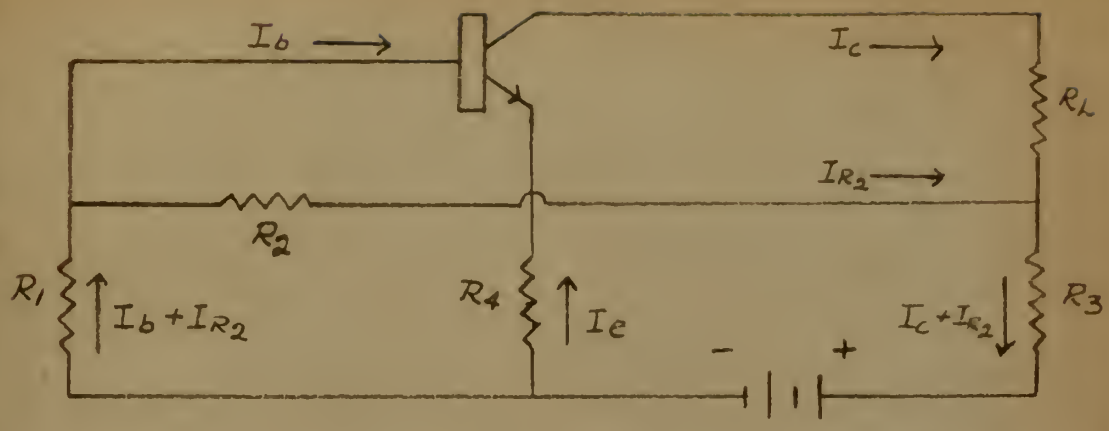
STABILIZATION AGAINST TEMPERATURE VARIATIONS

The circuit of Figure IV-1 (a) represents the d-c. circuit of Figure 36, Chapter V. The resistor R_L represents the d-c. resistance of the upper half of the primary winding of the transformer.

In Figure IV-1 (b), point O is the mean temperature biasing point. In this instance, point Q was established at 2.5 ma. collector current and 15 volts collector potential. The dashed line represents the d-c. load line. The d-c. load here is the 60 ohms resistance of the transformer winding mentioned above. Drawn more accurately for this specific case, this line would be essentially vertical. The solid lines indicate the a-c. load line. The a-c. load line is about 6000 ohms.

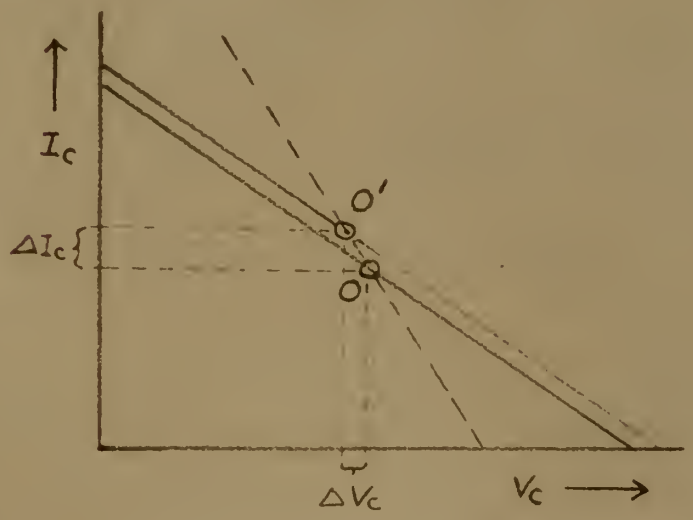
Any change in I_c, the d-c. collector current, will result in point O shifting to O'. An increase in the collector saturation current, I_{co}, as a result of increasing temperature produces exactly this kind of variation and to stabilize the output level, this shift must be confined within acceptable limits.

Since this is an oscillator circuit and the amplitude is limited essentially by the nonlinearities occurring at the points where the a-c. load line intersects the axes of the collector characteristics, the shift of the biasing point



R_2 may be the d-c. resistance of transformer primary

(a)



(b)

Figure IV

upward will permit the a-c. collector current, i_c, to vary in amplitude by an amount equal to ΔI_c. For the case with the d-c. load line nearly vertical, ΔV_c is essentially zero.

The first step in arriving at values for resistors R_1, R_2, R_3 and R_4 is to determine the maximum permissible ΔI_c.

It was decided to restrict variations in i_c to 0.5 db if possible. Then

$$0.5 = 20 \log \frac{2.5+\Delta I_c}{2.5}$$

From which $\Delta I_c = 0.265$ ma.

Next it is necessary to determine the variation in the saturation current, ΔI_{co}, this ΔI_c will allow. To accomplish this, the junction temperature is required, and to determine the junction temperature, the maximum power dissipation must be known.

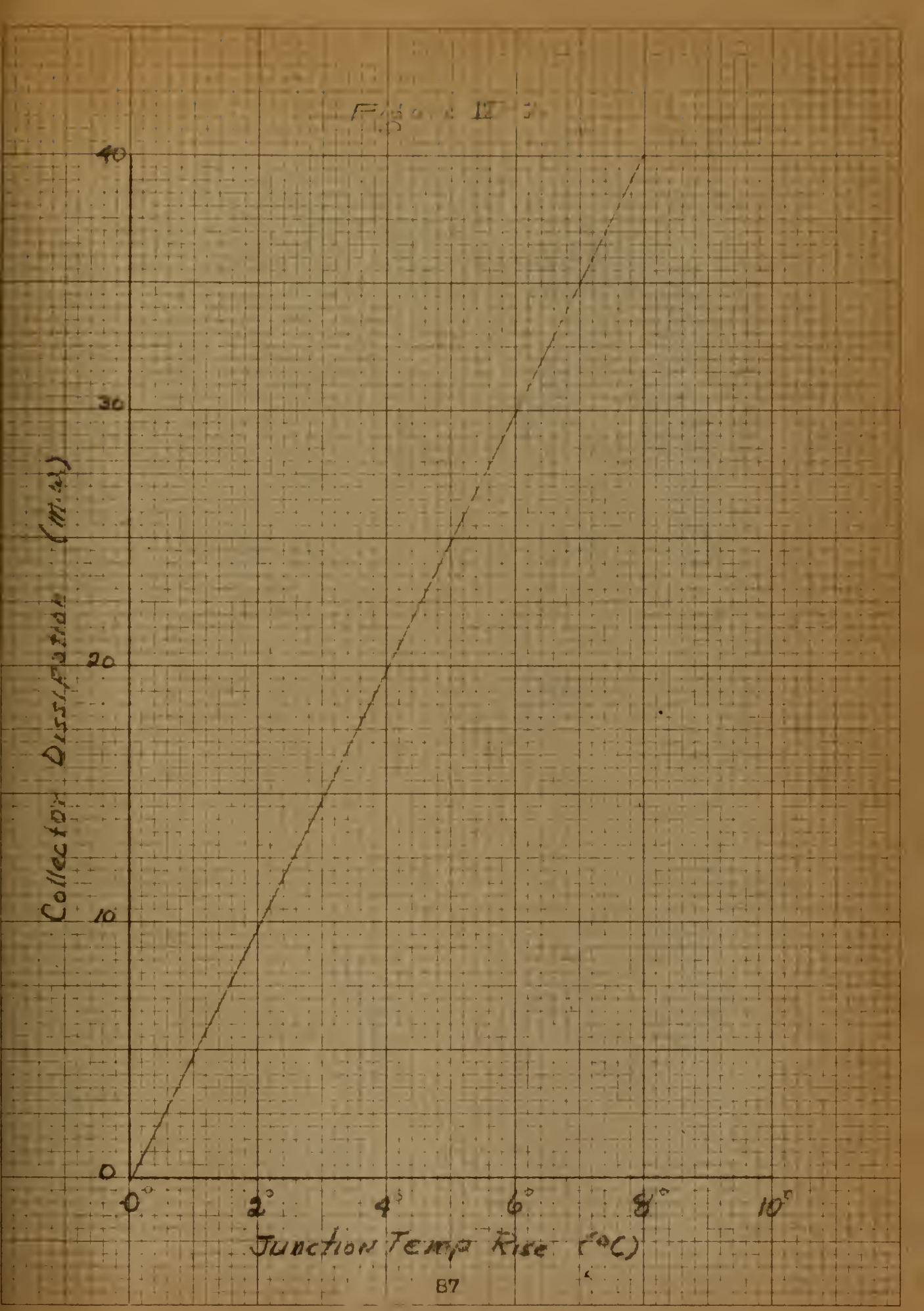
$$P_{max} = (I_c + \Delta I_c)(V_c - \Delta V_c)$$

$$P_{max} = (2.5 + 0.265)(10^{-3})(15)$$

with ΔV_c being considered negligible.

A linear extrapolation from Figure IV-2 indicates a rise of junction temperature of 8.3° centigrade for this dissipation. For an ambient temperature of 60° C., a junction temperature of 68.3° will exist.

For these computations, the average transistor was assumed to have an alpha of 0.935 and a collector saturation current of 6.0 microamperes at 25° C.


40
30
20
10
0
Collector Dissipation (mW)
0°
2°
4°
6°
8°
10°
Junction Temp Rise (°C)

Variations in saturation current with temperature are specified in Figure IV-3 to a close approximation for junction transistors such as that being used. For a junction temperature of 68.3°, the saturation current is seen to be 15.2 times the saturation current at 25°. Or,

$$I_{co(68.3°C)} = 15.2 \times 6.0 = 91.2\,\mu a.$$

Then,

$$\Delta I_{co} = 91.2 - 6.0 = 85.2\,\mu a.$$

$$P \equiv \frac{\Delta I_c}{\Delta I_{co}} = \frac{265}{85.2} = 3.1$$

Next,

$$R_2 = \frac{(P-1)V_{cc}}{\alpha(I_c - PI_{co})} = \frac{2.1 \times 130}{0.935(2.5 - 0.265)10^{-3}}$$

$$R_2 = 131.5 \text{ K-ohms.}$$

As shown in Figure IV-1 (a), the current through R_2,

$$I_{R_2} = \frac{V_c + I_c R_L}{R_2} = \frac{15 + (2.5 \times 10^{-3} \times 60)}{131.5 \times 10^3} = 0.115 ma.$$

Next, the equation

$$(I_{R_2} + I_c)R_3 = (V_{cc} - V_c - I_c R_L) - I_e R_4$$

may be written. This equation contains two unknowns, R_3 and R_4, and may be solved by arbitrarily assigning a value to one of them. Resistor R_4 is taken as 4.7 K-ohms.

The d-c. emitter current

$$I_e \approx \frac{I_c}{\alpha} = \frac{2.5}{0.935} = 2.675 ma.$$

Then,

$$R_3 = \frac{(V_{cc} - V_c - I_c R_L) - I_e R_4}{I_{R_2} + I_c}$$

$$I_{co}(T^\circ) \approx I_{co}(25^\circ)\, 2^{\left(\frac{T^\circ - 25^\circ}{11^\circ}\right)}$$

Figure III-3

$$R_3 = \frac{(130 - 15.15) - (2.675 \times 10^{-3} \times 4.7 \times 10^{3})}{2.615 \times 10^{-3}}$$

$$R_3 = 48.1 \text{ K-ohms.}$$

Last,

$$R_1 \approx \frac{I_e R_4}{I R_2 + I_b}$$

where the d-c. base current I_b is defined very closely by

$$I_b = I_e(1 - \alpha)$$

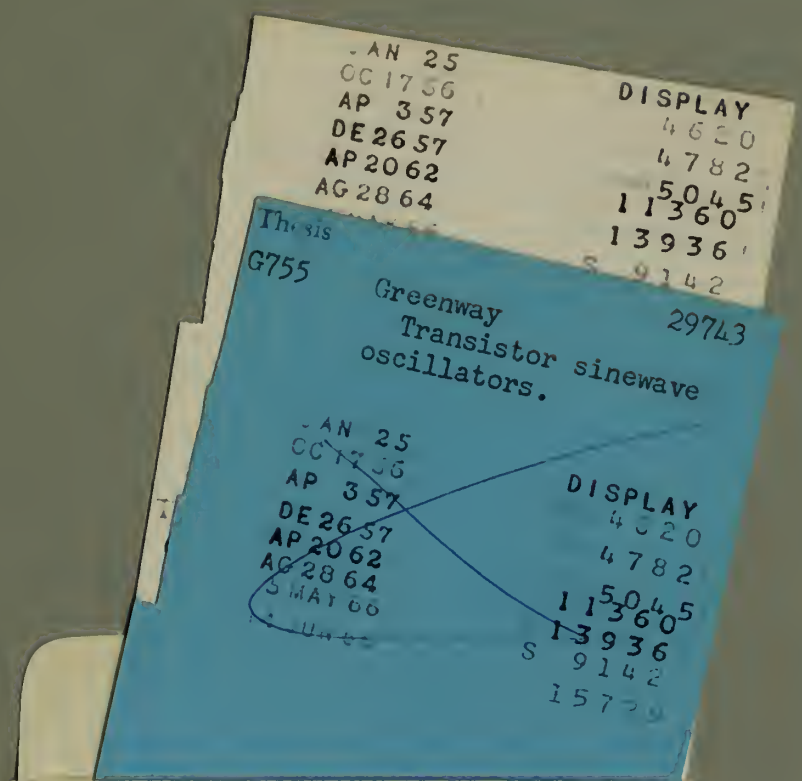
and in this case,

$$I_b = 2.675 \times 0.065 = 0.175 \text{ ma.}$$

$$R_1 \approx \frac{2.675 \times 10^{-3} \times 4.7 \times 10^{3}}{(0.115 + 0.175) \times 10^{-3}}$$

$$R_1 \approx 43.9 \text{ K-ohms.}$$

With the values of the resistors determined, the nearest RMA values were next selected and one-percent, deposited-carbon, high-stability resistors specified. These values are shown on page 66.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DE 26 57
AP 20 62
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